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Permanent Magnet Synchronous Generator Systems

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10.1 Introduction

By permanent magnet (PM) synchronous generators (SGs), we mean here radial or axial airgap PM brushless generators with distributed ($q > 1$) or concentrated ($q \leq 1$) windings and rectangular or sinusoidal current control with surface PM or interior PM (IPM) rotors. Multiple pole transverse flux machines (TFMs) or flux reversal machines (FRMs) conceived for low speed and high torque density will be discussed in [Chapter 11](#).

A PMSG's output voltage amplitude and frequency are proportional to speed. In constant speed prime-mover applications, PMSGs might perform voltage self-regulation by proper design; that is, inset or interior PM pole rotors. Small speed variation (± 10 to 15%) may be acceptable for diode rectified loads with series capacitors and voltage self-regulation. However, most applications require operation at variable speed, and, in this case, constant output voltage vs. load, be it direct current (DC) or alternating current (AC), requires full static power conversion and close-loop control.

Versatile mobile generator sets (gensets) use variable speed for fuel savings, and PMSGs with full power electronics control can provide high torque density, low losses, and multiple outputs (DC and AC at 50 [60] Hz or 400 Hz, single phase or three phase).

A high efficiency, high active power to peak kilovoltampere (kVA) ratio allows for reasonable power converter costs that offset the additional costs of PMs in contrast to switched reluctance generators (SRGs) or induction generators (IGs) for the same speed.

For automotive applications, and when motoring is not necessary, PM generators may provide controlled DC output for a 10 to 1 speed range through a diode rectifier and a one insulated gate bipolar transistor (IGBT) step-up DC–DC converter for powers above 2 to 3 kW. A series hybrid vehicle is a typical application here. Gas turbines run at super high speeds; 3.0 megawatt (MW) at 18 krpm to 150 kW at 80 krpm. Direct-driven super-high-speed PM generators, with their high efficiency and high power factor, seem to be the solution for such applications. With start-up facilities for bidirectional power flow, static converters allow for four-quadrant control at variable speed, with $\pm 100\%$ active and reactive power capabilities. Distributed power systems of the future should take advantage of this technology of high efficiency, reasonable cost, and high flexibility in energy conversion and in power quality.

Flywheel batteries with high kilowatts per kilogram (kW/kg), good kilowatthours per kilogram (kWh/kg), and long life, also make use of super-high-speed PMSGs with four-quadrant P and Q control. They are proposed for energy storage on vehicles and spacecraft and for power systems backup.

Diverse as they may seem, these applications are accommodated by only a few practical PMSGs classified as follows:

- With radial airgap (cylindrical rotor)
- With axial airgap (disk rotor)
- With distributed stator windings ($q > 1$)
- With concentrated windings ($q \leq 1$)
- With surface PM rotors
- With interior or inset PM rotors
- With rectangular current control
- With sinusoidal current control

In terms of loads, they are classified as follows:

- With passive AC load
- With DC load
- With controlled AC voltage and frequency at variable speed

The super-high-speed PM generators differ in rotor construction, which needs a mechanical shell against centrifugal forces and a copper shield (damper) to reduce rotor losses. Also, at high fundamental frequency (above 1 kHz), stator skin effect and control imply special solutions to reduce machine and static converter losses and overall costs.

As in most PMSG surfaces PM rotors are used, the latter will be given the most attention. An IPM rotor case will be covered in a single paragraph, when voltage self-regulation is acceptable due to almost constant speed operation. Basic configurations for stator and rotor will be introduced and characterized. A comprehensive analytical field model is introduced and checked through finite element method (FEM) field and torque production analysis. Loss models for generator steady-state circuit modeling are introduced for rectangular and for sinusoidal current control. Design issues and a methodology by example are treated in some detail.

PM generator control and its performance with direct AC loads, with rectified loads, and with constant voltage and frequency output at variable speed by four-quadrant AC-AC static power converter P - Q control systems are all treated in separate sections. Super-high-speed PM generator design and control are dealt with in some detail, with design issues as the focal point. Methods for testing the PMSGs to determine losses, efficiency, and parameters close the present chapter.

10.2 Practical Configurations and Their Characterization

PM brushless motor drives have become a rather mature technology with a sizeable market niche worldwide. Thorough updates of these technologies may be found in the literature [1–5].

For PMSGs, both surface PM and inset PM pole rotors are used. A typical cylindrical rotor configuration is shown in Figure 10.1. For IPM pole rotors, the magnetic reluctance along the direct (d) axis is larger than for the transverse (q) axis; thus, $L_d < L_q$ — that is, inverse saliency, in contrast to electromagnetically excited pole rotors for standard synchronous machines.

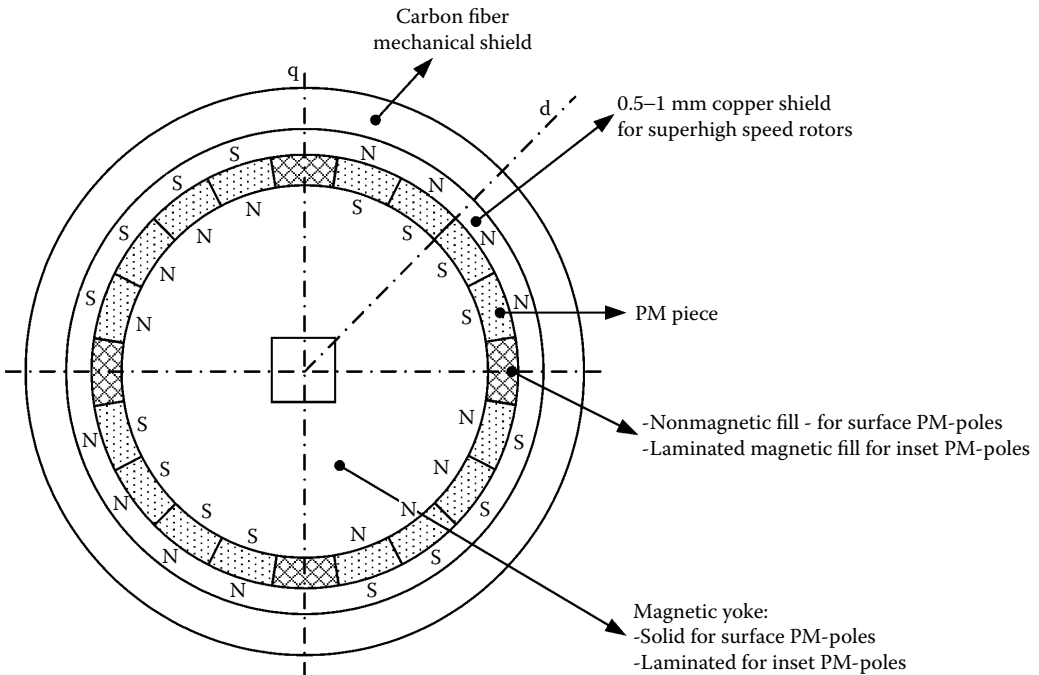


FIGURE 10.1 Four-pole surface permanent magnet (PM) and inset PM pole rotor configurations.

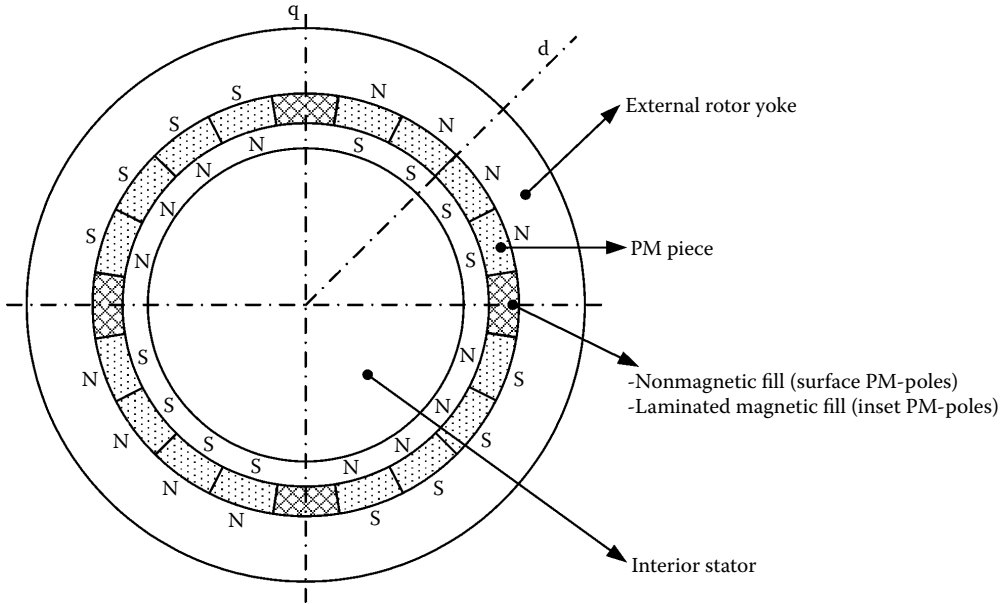


FIGURE 10.2 Four-pole cylindrical external rotor.

The d axis falls along the PM field axis in the airgap. The rotor may be internal (Figure 10.1) or external (Figure 10.2) to the stator, in cylindrical rotor configurations. Interior rotors require a carbon fiber mechanical shield (retainer) against centrifugal forces for high-speed applications (above 50 to 80 m/sec peripheral speed). In contrast, external rotors do not need such a retainer, but the yoke has to withstand high centrifugal forces in high-speed rotors.

For high-speed PMSGs with cylindrical rotors, characterized by a wide constant power speed range ($\omega_{max}/\omega_b > 4$), a hybrid surface PM pole and variable reluctance rotor may be used (Figure 10.3a through Figure 10.3c). The surface PM pole rotors were proven capable of slightly more torque for given rotor external diameter and length and the same PM weight. For generators, they provide for small $l_d = l_q$ inductances in per unit (P.U.) terms, which should result in lower voltage regulation.

However, inset PM pole or hybrid rotors, due to inverse saliency ($L_d < L_q$) may produce zero voltage regulation at a certain load that is fixed by design. In self-regulated PMSGs, driven at quasi-constant speed, such a design may prove to be practical.

In super-high-speed PMSGs with fundamental frequency above 1 kHz, the core loss and rotor loss are major concerns. To reduce rotor losses, the surface PM rotors are the natural choice. In such applications, copper shields surrounding the PM pole reduce the harmonics losses in the rotor PMs, and, moreover, solid back iron in the rotor is acceptable. This, in turn, leads to better rotor mechanical integrity.

Disk-shaped PM rotors were also proposed to increase the torque/volume in axial-length constraint designs. The typical rotor has the PMs embedded in a stainless disk; a mechanically resilient magnetic disk behind the PMs is also feasible, as it reduces the rotor harmonics losses (Figure 10.4a and Figure 10.4b). The configurations in Figure 10.4 preserve the ideal zero axial force on the central part when the latter rotor (stator) is in the center position; thus, the bearings are spared large axial forces.

It is, in principle, possible to use only one of the two rotor parts in Figure 10.4b with a single stator in front of it, axially. But in this case, the axial (attraction) force between rotor and stator is very large, and the bearings have to handle it. Special measures to reduce vibration and noise are required in this situation.

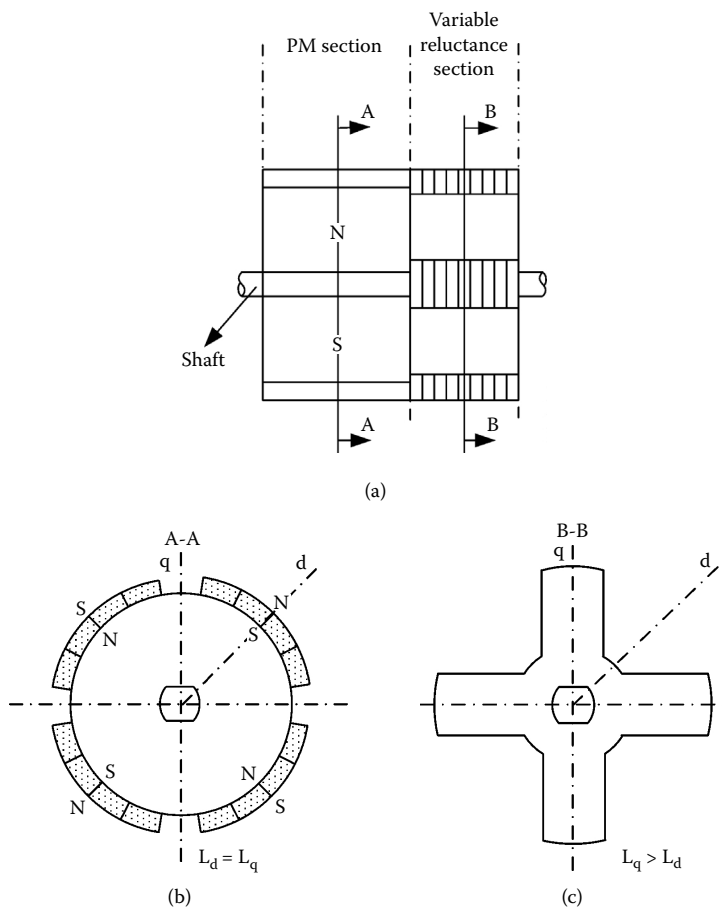


FIGURE 10.3 Four-pole permanent magnet (PM) rotor: (a) with surface PM pole, (b) A-A cross-section, and variable reluctance, (c) B-B cross section.

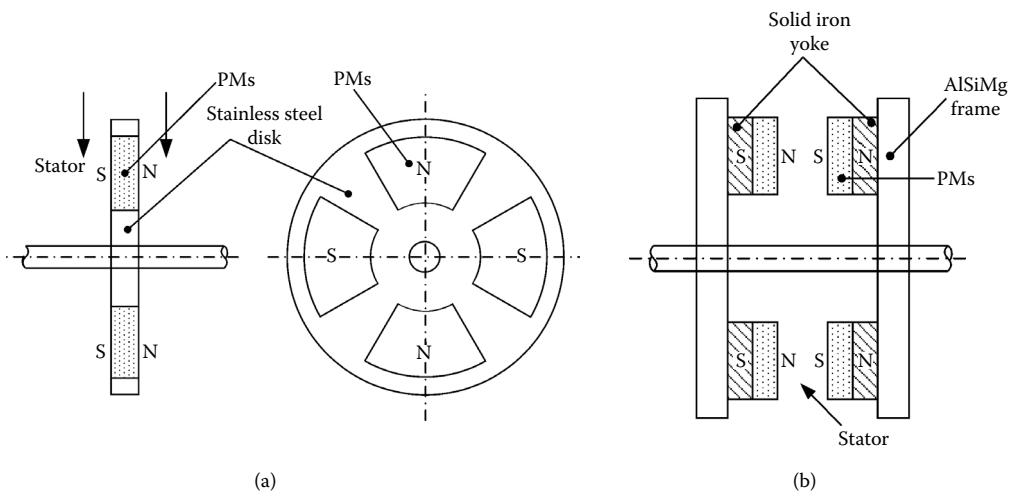


FIGURE 10.4 Four-pole disk-shape permanent magnet (PM) rotors: (a) interior rotor (with dual stator) and (b) double-sided rotor (with single and dual face stators).

In the interior rotor configuration, there is no rotor magnetic core in Figure 10.4a, and the axial rotor length and its inertia are small. In contrast, the two external stators will require a magnetic yoke (each of them). To reduce the stator's axial length, the number of poles must be increased. This renders the disk rotor PM generators favorable for a large number of poles ($2p_1 > 6$) for fundamental frequency below 2 kHz. Multiple rotor and stator disk rotor configurations were proposed for low-speed high-torque direct drives for elevators [5].

Though quite a few thorough comparisons between cylindrical and disk rotor PM synchronous machines were performed, no clear-cut "prescriptions" are available. However, it seems that depending on the number of poles ($2p_1 > 8, 10$) and if the stack length is short, the disk rotor PM synchronous generators lead to higher torque/volume for equivalent losses and torque [6, 7].

For large torque (diameter) applications, such as direct-driven wind generators, low-cost ferrite PMs were proposed to cut the costs [8]. Heavy PM flux concentration is required. A typical such rotor structure is shown in Figure 10.5.

Apparently, the configuration exhibits saliency ($L_d < L_q$) due to the presence of the PMs along the d axis field path. In reality, the slim flux bridges (h_b) lead to their early saturation by the q axis (torque) currents in the stator; thus, the saliency decreases with load to negligible values. A small machine inductance generally produces small voltage regulation.

To reduce the tangential fringing flux of the PMs in the airgap, $b_{PM} > 3$ to $4g$ (g is the mechanical gap between rotor and stator). For rotor diameters of 3 to 4 m, even at speeds below 30 rpm, typical for large wind turbines, a mechanical airgap of at least 5 mm is mechanically required. Therefore, the PM width b_{PM} should be at least 15 to 20 mm, and the pole pitch τ should not be less than 80 to 100 mm to provide smooth PM flux density distribution in the airgap.

The silicon iron laminated poles of the rotor lead to moderate rotor surface and pulsation harmonics core losses in the rotor.

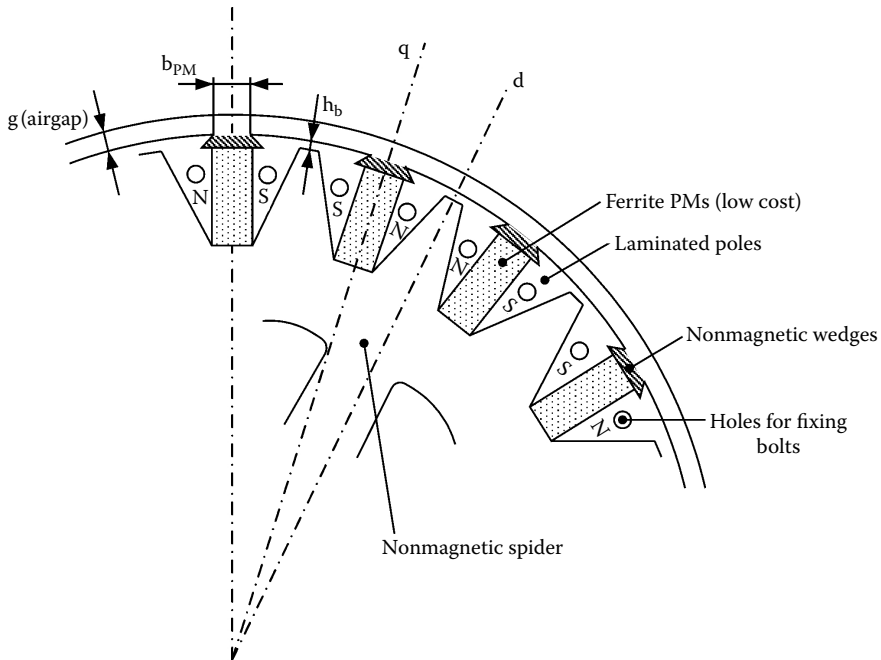


FIGURE 10.5 Ferrite permanent magnet (PM) rotor with flux concentration.

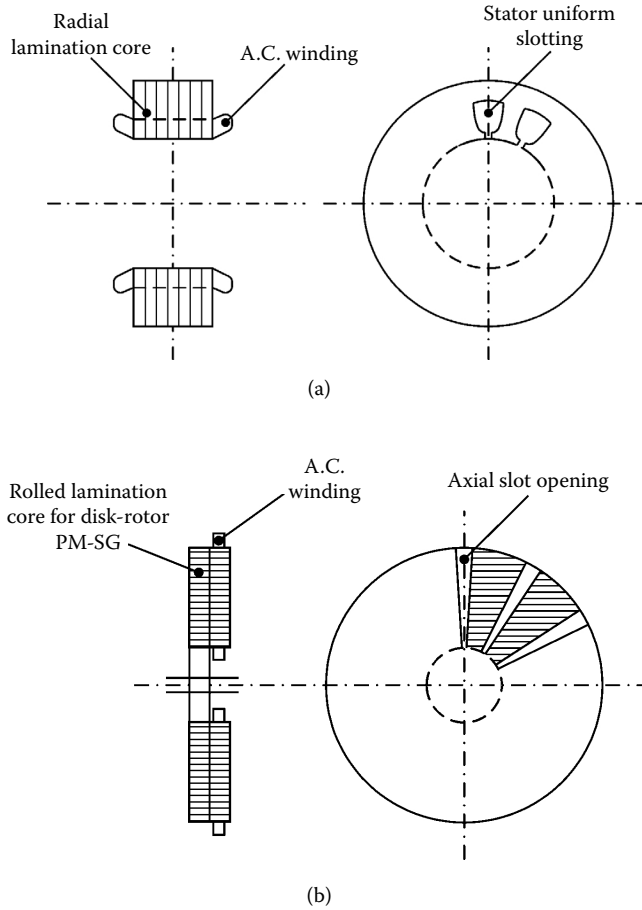


FIGURE 10.6 Slotted stator cores: (a) for cylindrical-rotor permanent magnet synchronous generator (PMSG) and (b) for disk-rotor PMSG.

Generally, stators are made of a laminated iron core with slots, but slotless cores may be adopted for high-speed (frequency) applications. The core may be cylindrical or, again, disk shaped (Figure 10.6a and Figure 10.6b).

The radial and axial slots are stamped into, respectively, radial and rolled silicon-iron lamination cores.

In super-high-speed applications, the core losses in the stator teeth may become so large that their elimination could prove beneficial, despite PM flux density reduction due to increased “magnetic” airgap. Slotless stators for cylindrical and disk-shaped rotor PMSGs are shown in Figure 10.7a and Figure 10.7b.

The space left between Gramme-ring coils is required to fix the rolled lamination stator core of the disk rotor PMSG (Figure 10.7b) to the machine frame. Still, the stator structure is a bit mechanically fragile, and care must be exercised in its mechanical design to avoid inadmissible axial bending.

The Gramme-ring winding is a special winding with short end connections, but a standard three-phase winding may be placed, as for the cylindrical rotor machine, in a dummy, isolation-made, slotted winding holder to reduce manufacturing time and costs.

As expected, being “washed” by the cooling air, the stator slotless windings may be better cooled, but, due to the larger magnetic airgap, the winding loss and torque tend to be notably larger than that in slotted cylindrical stator configurations.

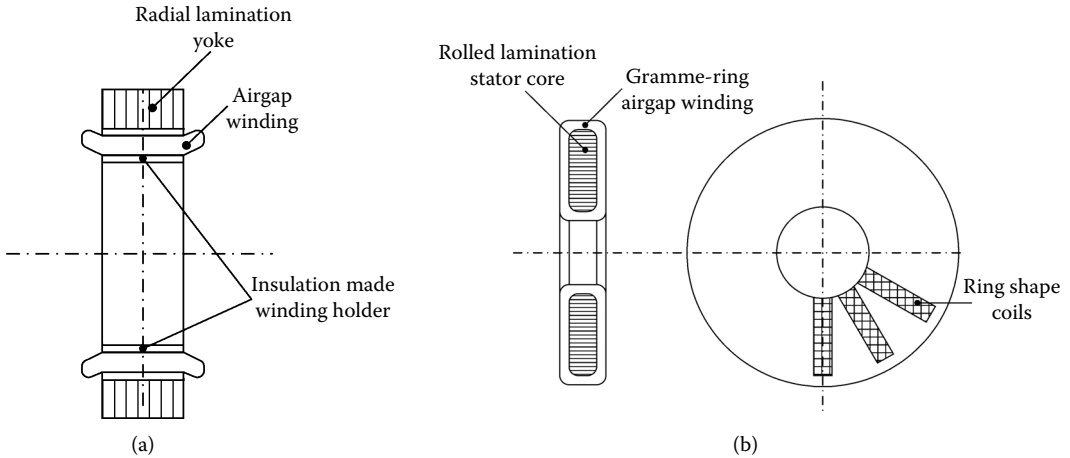


FIGURE 10.7 Airgap windings: (a) for cylindrical rotor and (b) for double-sided disk-shaped rotor.

10.2.1 Distributed vs. Concentrated Windings

Distributed single- or double-layer three-phase windings with $q = 2 \div 6$ are typical for PMSGs. Chording of coils in a ratio of about 5:6 is performed on double-layer windings to reduce the fifth and seventh stator magnetomotive force (mmf) space harmonics and their additional rotor surface and PM eddy current losses.

A large q (slot/pole/phase) leads to an increase in the order of the first slot harmonic, with amplitude that is thus reduced along with its additional rotor surface and PM eddy current losses. However, for large q , the end connections tend to be large, and thus, the Joule losses tend to increase. Up to $q = 3$, a good compromise in terms of total losses may be secured.

Distributed windings are used to provide an almost sinusoidal electromagnetic field (emf) in the stator winding despite the nonsinusoidal (rather flat) PM flux density distribution in the airgap. Stator or rotor skewing may be added to further sinusoidalize the emf at the price of a 5 to 7% reduction in emf root mean squared (RMS) value for one stator slot pitch skewing. Sinusoidal emf requires sinusoidal shape current to provide, ideally, rippleless interaction (PM to winding currents) torque.

The rather trapezoidal emf is obtained with three slots per pole ($q = 1$). If the PM airgap flux density distribution is flat in this case, trapezoidal distribution (ideally rectangular) current control is required to reduce torque pulsations between the commutation of phases. For rectified output PMSGs, the exploitation of the third harmonic in the emf can lead to increased power conversion; so trapezoidal emf may be adopted in some designs.

In an effort to reduce stator manufacturing costs and the winding Joule losses, concentrated (non-overlapping) windings with $q < 1$ were adopted [9, 10].

The number of stator slots N_s and the number of rotor PM poles $2p_1$ are close to each other. The closer to unity is the ratio $N_s/2p_1$, the better, as the number of periods of fundamental torque at zero current due to PMs and slot openings is equal to the smallest common multiplier (SCM) of N_s and $2p_1$. Also, the equivalent winding factor for the winding is large if the SCM is large.

The nonoverlapping coil windings may be built in one layer or two layers (Figure 10.8a through Figure 10.8c). The single-layer concentrated windings (Figure 10.8b) lead to only slightly higher winding factors (Table 10.1), but they imply longer end connections and, thus, longer stator frames in comparison to double-layer concentrated windings. The latter allow for a slightly less slot filling factor, as there are two coils in a slot and, inevitably, some airspace remains between the two coils [10].

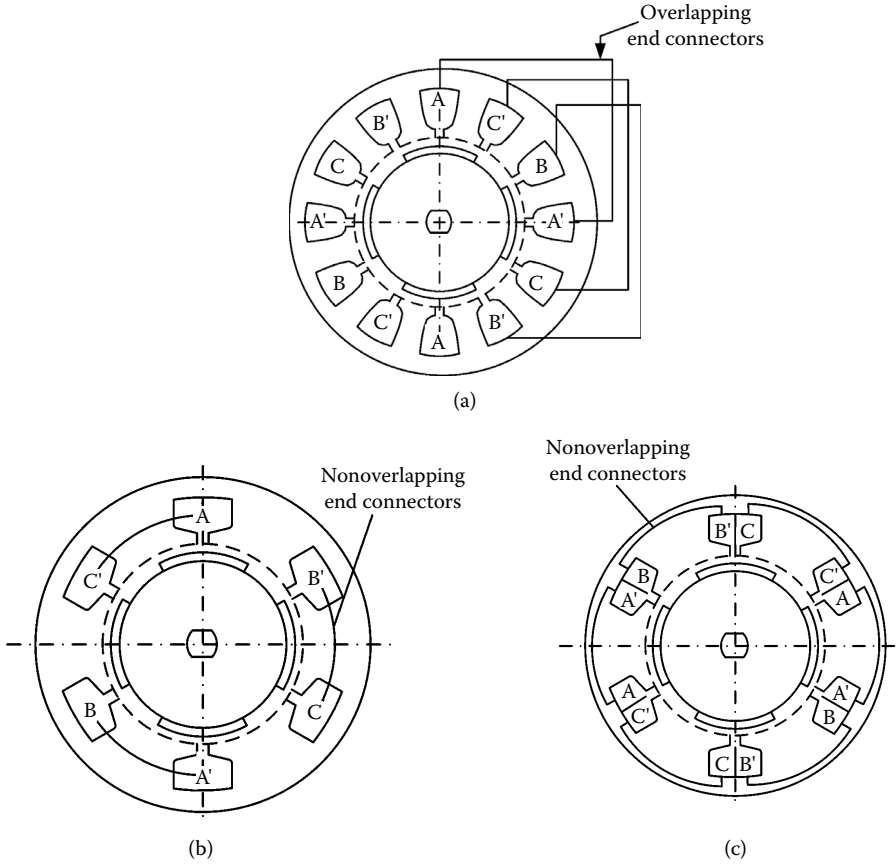


FIGURE 10.8 Four-pole winding layouts for permanent magnet synchronous generators (PMSGs): (a) distributed winding ($q = 1$); (b) single-layer concentrated winding; and (c) double-layer concentrated winding.

The winding factor is defined as the ratio of the resulting emf E_{cp} per current path (or phase) divided by the product of the number of coils N_{cp} to their emfs E_c :

$$k_w = \frac{E_{cp}}{N_{cp} \cdot E_c} \quad (10.1)$$

Values of k_w 0.866 compare favorably with distributed windings. There are some side effects, such as the presence of both odd and even harmonics in the stator three-phase mmf. As such, harmonics count as leakage flux if they lead to an increase in the leakage inductance per current path (or phase). Some single-layer concentrated windings produce subharmonics that create additional core losses and further leakage inductance increases. This is not so, in general, for double-layer concentrated windings [10].

The cogging torque number of periods is given by the SCM of the N_s and $2p_1$. Some representative data are given in Table 10.2. As can be seen by comparing Table 10.1 and Table 10.2, the maximum SCM does not always coincide with the maximum winding factor; thus, a compromise between the two is required, with the latter being more important. For $N_s/2p_1 = 15/14, 8/9, 12/10, 12/14, 18/16$, very good results are obtained.

TABLE 10.1 Winding Factors of Concentrated Windings

	$2p$ — Poles															
	2		4		6		8		10		12		14		16	
N_s Slots	*	**	*	**	*	**	*	**	*	**	*	**	*	**	*	**
3	*	0.86	*	*	*	*	*	*	*	*	*	*	*	*	*	*
6	*	*	0.866	0.866	*	*	0.866	0.866	0.5	0.5	*	*	*	*	*	*
9	*	*	0.736	0.617	0.667	0.866	0.960	0.945	0.96	0.945	0.667	0.764	0.218	0.473	0.177	0.175
12	*	*	*	*	*	*	0.866	0.866	0.966	0.933	*	*	0.966	0.933	0.866	0.866
15	*	*	*	*	0.247	0.481	0.383	0.621	0.866	0.866	0.808	0.906	0.957	0.951	0.957	0.951
18	*	*	*	*	*	*	0.473	0.543	0.676	0.647	0.866	0.866	0.844	0.902	0.960	0.931
21	*	*	*	*	*	*	0.248	0.468	0.397	0.565	0.622	0.521	0.866	0.866	0.793	0.851
24	*	*	*	*	*	*	*	*	0.930	0.463	*	*	0.561	0.76	0.866	0.866

Note: * = one layer; ** = two layers.

Source: Adapted from F. Magnussen, and C. Sadarangani, Winding factors and Joule losses of PM machines with concentrated windings, Record of IEEE-IEMDC-2003, vol. 1, 2003, pp. 333–339.

TABLE 10.2 Smallest Common Multiplier (SCM) of N_s and $2p_1$

N_s	$2p_1$							
	2	4	6	8	10	12	14	16
3	6	12	*	*	*	*	*	*
6	*	12	*	24	30	*	*	*
9	*	36	18	72	90	36	126	144
12	*	*	*	24	60	*	84	48
15	*	*	30	120	30	60	210	240
18	*	*	*	72	90	36	126	144
21	*	*	*	168	210	84	42	336
24	*	*	*	*	120	*	168	48

A large SCM means that, in general, low cogging (zero current) torque is obtained without skewing. A few percent (5 to 7%) output is saved this way.

As the number of poles $2p_1$ is increased, with $N_s \approx 2p_1$, both the winding factor and the SCM increase, securing good performance. This means that for some low-speed applications, concentrated windings are definitely favorable.

By adequate rotor geometry, the emf may be made quasi-sinusoidal, and thus, vector and direct torque control are feasible. However, when the airgap has to increase due to mechanical reasons (higher diameter, etc.), the tangential fringing flux of the PMs tends to increase, reducing the emf. Poor utilization of PMs takes place, and reduced output is the result. When the slots are thin (in small torque machines), there may not be enough mmf per slot to secure high torque density. With these limitations in mind, concentrated windings may represent a practical solution to high-performance PMSGs in various applications.

The Gramme-ring windings in Figure 10.7 may be assimilated to a single-layer concentrated or distributed winding by making adequate connections of coils.

Note that so far, we considered the stator cores as being made of single, stamped laminations. For outer stator diameter above 1 m, the stator core has to be made from a few sections, as is usually the case with SGs (Chapter 5). Modular stator cores are plagued with noncircularity eccentricities that create additional losses, torque pulsations, vibration, and noise.

10.3 Airgap Field Distribution, emf and Torque

The airgap field in PMSGs has two main sources: the PMs and the stator mmf.

The presence of slot openings and magnetic saturation, together with the rotor PM geometry, makes the computation of airgap field in PMSG a complex problem, solvable as it is only by two-dimensional (2D) or three-dimensional (3D) finite element method (FEM).

However, at least for surface PM pole rotors, the influence of magnetic saturation may be neglected unless very high current loading is allowed for, in very high torque density designs. Once magnetic saturation is neglected, the superposition of PM and stator mmf fields in the airgap is acceptable.

Moreover, the effect of stator slot openings on airgap field distribution may be added by introducing a P.U. (relative) airgap performance function $\lambda(\theta)$, variable with angular position along the stator surface (Figure 10.9). Consequently, a 2D analytical field model may be used to calculate separately the PM and stator mmf produced airgap field distribution in the slotless machine, and then the slot opening effect can be introduced by multiplication with $\lambda_{P.U.}(\theta)$:

$$B_{PM,mmf}(\theta) = B_{PM,mmf}^{slotless}(\theta - \theta_1) \cdot \lambda_{P.U.}(\theta) \quad (10.2)$$

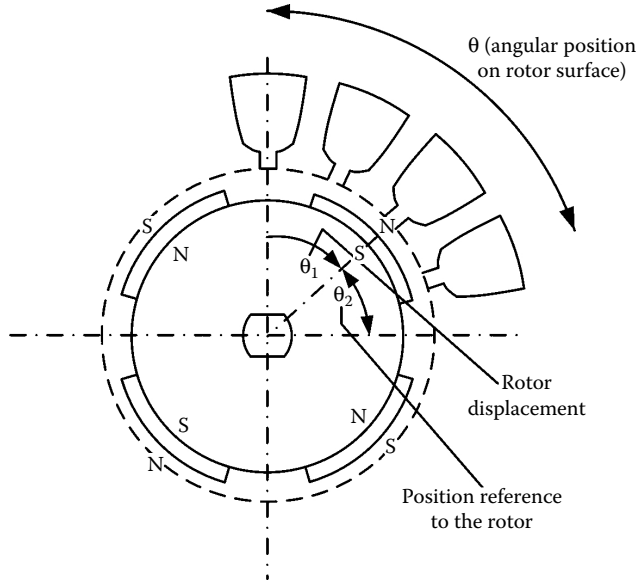


FIGURE 10.9 Position angles with rotor in motion.

where θ_1 is the rotor displacement, and θ and θ_1 are taken in electrical degrees or radians. Complete 2D analytical models could be built for the unsaturated PMSG with surface PM pole cylindrical rotors.

For the axial airgap (disk-shaped rotors), a similar 2D theory in the axial (z) and in polar coordinates (θ_1, θ) may be developed, while the variation of variables in the radial direction, r , is handled by an additional P.U. coefficient $\lambda_{p.u.}(r)$. A quasi-3D analytical model is thus obtained. A complete 2D analytical model for a cylindrical rotor is introduced in Reference [11] and pursued further in Reference [12] for cogging torque and emf computation. For the disk-shaped rotor case, a quasi-3D complete analytical model is introduced in References 13 and 14 with emf and cogging torque thoroughly documented. In both cases, the PM equivalent mmf and the stator mmf distributions are decomposed in Fourier space harmonics, and superposition is applied afterward.

As these analytical models are fairly general, and could handle both distributed and concentrated stator windings, we will treat these here in some detail in view of their potential application to design optimization attempts.

The equivalent circuit for phase a in stator coordinates is shown in Figure 10.10. Core losses are neglected. E_{PMa} represents the PM-induced voltage (emf) in phase a which, evidently, is a bipolar motion-induced voltage that varies with the rotor position:

$$E_{PMa} = \sum_{i=1}^{N_s/3} W_c \cdot B_{PMi} \cdot l_{stack} \cdot u \quad (10.3)$$

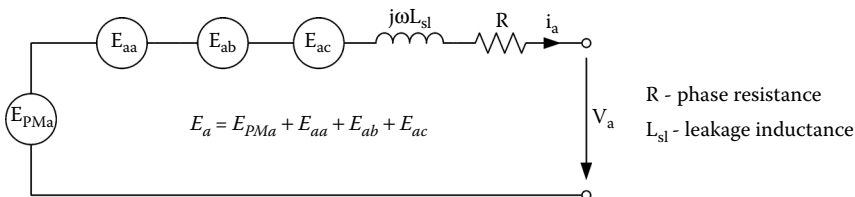


FIGURE 10.10 Steady-state phase equivalent circuit.

where

l_{stack} is the stack

u is the peripheral speed

B_{PMi} is the PM average flux density at slot i of phase a

W_c is the number of turns per coil

N_s is the stator slots

E_{aa} , E_{ab} , and E_{ac} are the self-induced and mutual-motion-induced voltages in phase a . They are produced by the stator currents:

$$\begin{aligned} E_{aa} &= \sum_{i=1}^{N_s/3} \left(\sum_{j=1}^{N_c} W_c \cdot B_{cij}^a \cdot l_{stack} \cdot u \right) \\ E_{ab} &= \sum_{i=1}^{N_s/3} \left(\sum_{j=1}^{N_c} W_c \cdot B_{bij}^a \cdot l_{stack} \cdot u \right) \\ E_{ac} &= \sum_{i=1}^{N_s/3} \left(\sum_{j=1}^{N_c} W_c \cdot B_{cij}^a \cdot l_{stack} \cdot u \right) \end{aligned} \quad (10.4)$$

where B_{aij}^a is the average instantaneous magnetic flux density produced by coil i of phase a at slot j of the same phase a . B_{bij}^a , B_{cij}^a are the average instantaneous magnetic flux densities produced by coil i of phase $b(c)$ at slot j of phase a . That is, mutual stator emfs E_{aa} , E_{ab} , E_{ac} vary with rotor position if inset or interior (or variable reluctance) rotor poles exist.

When considering the following, the approximations in Figure 10.11a through Figure 10.11d are operated:

- PM magnetization harmonics
- Current sheet distribution of a coil
- Stator slot openings

The rectangular PM magnetization (Figure 10.10) may be decomposed in Fourier series:

$$M(\theta_2) = \sum_{n=1,3,5}^{\infty} 2 \frac{B_r}{\mu_0} \cdot \alpha_p \cdot \frac{\sin(n\pi\alpha_p/2)}{n\pi\alpha_p/2} \cos np_1\theta_2 \quad (10.5)$$

where

$\alpha_p = \tau_m/\tau$ is the magnet/pole pitch ratio

p_1 is the pole pairs

B_r is the remnant flux density of the PMs

$\theta_2 = \theta - \theta_1$

The stator current sheet produced by a coil (Figure 10.11b) may also be approximated in Fourier series:

$$I(\theta) = \frac{4W_c i}{\pi w_s} \sum_{n=1}^{\infty} \frac{1}{n} \cdot \sin \frac{nw_s}{2R_s} \cdot \cos(n \cdot (\theta - \tau_c/2)) \quad (10.6)$$

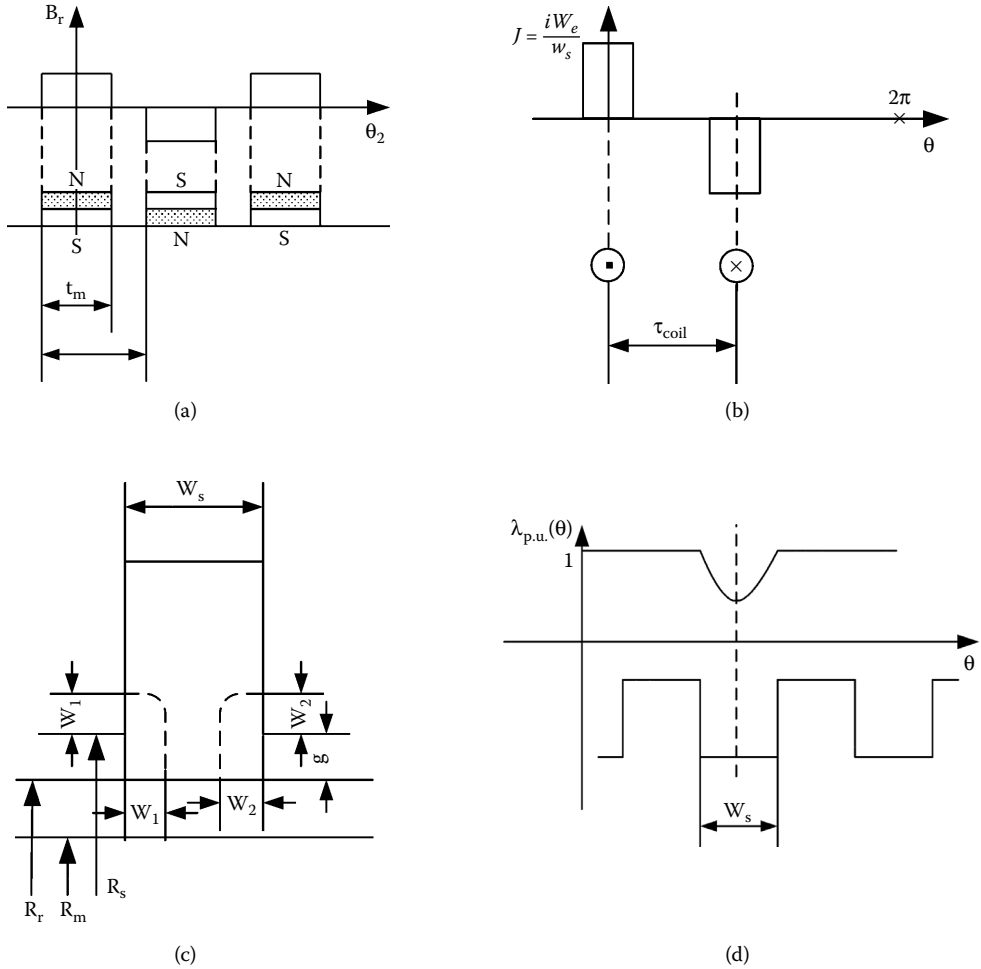


FIGURE 10.11 (a) Permanent magnet (PM) magnetization, (b) current sheet, (c) slot flux line ideal distributions, and (d) $\lambda_{p.u.}(\theta)$.

where

i is the instantaneous current in the coil

w_s is the slot opening

R_s is the interior stator radius

τ_c is the coil span angle

The relative airgap permeance (Figure 10.11c) $\lambda_{p.u.}(\theta)$ may be determined by considering semicircular flux paths in the slots with radii equal to the shortest distance to tooth edges:

$$\lambda_{p.u.}(\theta) = \frac{g + \frac{l_m}{\mu_r}}{g + \frac{l_m}{\mu_r} + \frac{\pi}{2} w_1(\theta)} \quad (10.7)$$

Away from slot openings, $\lambda_{p.u.}$ is unity and, in front of slot openings, has a variable subunity value, dependent on w_1 , which is the distance to the closest slot edge. For semiclosed slots, w_s refers to the slot opening of the airgap and not to the actual slot width, and thus, the airgap permeance pulsations with θ are notably smaller, as expected.

Making use of Equation 10.2, the expressions of $B_{PM}^{slotless}(\theta - \theta_1)$ and $B_{mmf}^{slotless}(\theta - \theta_1)$ are obtained with Equation 10.5 and Equation 10.6 and the 2D field model [11]:

$$B_{PM}^{slotless}(\theta - \theta_1) = \sum_{n=1,3,5}^{\infty} 4 \cdot \frac{B_r}{\mu_r} \cdot \frac{\sin \frac{n\pi\alpha_p}{2}}{\pi\alpha_p} \cdot \frac{p_1}{(np_1)^2 - 1} \cdot \left[\frac{(np_1 - 1) + 2\left(\frac{R_r}{R_m}\right)^{np_1+1} - (np_1 + 1) \cdot \left(\frac{R_r}{R_m}\right)^{2np_1}}{\frac{\mu_r+1}{\mu_r} \cdot \left(1 - \left(\frac{R_r}{R_s}\right)^{2np_1}\right) - \frac{\mu_r-1}{\mu_r} \left(\left(\frac{R_m}{R_s}\right)^{2np_1} - \left(\frac{R_r}{R_m}\right)^{2np_1}\right)} \right] \cdot \left[\left(\frac{r}{R_s}\right)^{np_1-1} \cdot \left(\frac{R_m}{R_s}\right)^{np_1+1} + \left(\frac{R_m}{r}\right)^{np_1+1} \right] \cos np_1(\theta - \theta_1) \quad \text{for } p_1 \neq 1 \quad (10.8)$$

where μ_r is the P.U. value of PM permeability: 1.03 to 1.08. For $np_1 = 1$, Equation 10.8 has a zero denominator and should be changed to the following [11]:

$$B_{PM}^{slotless}(\theta - \theta_1) = \sum_{n=1,3,5}^{\infty} 2 \cdot \frac{B_r}{\mu_r} \cdot \frac{\sin \frac{n\pi\alpha_p}{2}}{n\pi\alpha_p} \cdot \frac{p_1}{(np_1)^2 - 1} \cdot \left[\frac{\left(\frac{R_m}{R_s}\right)^2 - \left(\frac{R_r}{R_s}\right)^2 + \left(\frac{R_r}{R_s}\right)^2 \ln \left(\frac{R_m}{R_r}\right)^2}{\frac{\mu_r+1}{\mu_r} \cdot \left(1 - \left(\frac{R_r}{R_s}\right)^2\right) - \frac{\mu_r-1}{\mu_r} \left(\left(\frac{R_m}{R_s}\right)^2 - \left(\frac{R_r}{R_m}\right)^2\right)} \right] \cdot \left(1 + \frac{R_s}{r}\right) \cos(\theta - \theta_1) \quad (10.9)$$

For the interior rotor,

$$R_m = R_s - g; \quad R_r = R_s - g - h_m \quad (10.10)$$

while for the exterior rotor,

$$R_m = R_s + g; \quad R_r = R_s + g + h_m \quad (10.11)$$

Consequently, Equation 10.8 and Equation 10.9 have to be modified, as now $R_s < R_r < R_m$:

$$B_{PM}^{slotless}(\theta - \theta_1) = - \sum_{n=1,3,5}^{\infty} 4 \cdot \frac{B_r}{\mu_r} \cdot \frac{\sin \frac{n\pi\alpha_p}{2}}{\pi\alpha_p} \cdot \frac{p_1}{(np_1)^2 - 1} \cdot \left[\frac{(np_1 - 1) \left(\frac{R_m}{R_r}\right)^{2np_1} + 2 \left(\frac{R_m}{R_r}\right)^{np_1-1} - (np_1 + 1)}{\frac{\mu_r+1}{\mu_r} \cdot \left(1 - \left(\frac{R_s}{R_r}\right)^{2np_1}\right) - \frac{\mu_r-1}{\mu_r} \left(\left(\frac{R_s}{R_m}\right)^{2np_1} - \left(\frac{R_r}{R_r}\right)^{2np_1}\right)} \right] \cdot \left[\left(\frac{r}{R_m}\right)^{np_1-1} + \left(\frac{R_s}{R_m}\right)^{np_1-1} \cdot \left(\frac{R_s}{R_r}\right)^{np_1+1} \right] \cos np_1(\theta - \theta_1) \quad \text{for } np_1 \neq 1 \quad (10.12)$$

The above expressions account for the rotor curvature and should be particularly instrumental for low rotor diameter (low torque) PMSGs.

The armature reaction flux density in the airgap, produced by a single stator coil with its sides placed in slots k and l , is as follows [12]:

$$B_{kl}(\theta) \equiv \frac{2\mu_0 W_c i}{\pi} \cdot \sum_{n=1}^{\infty} \frac{1}{r} \cdot \frac{\sin\left(n \frac{w_s}{2R_s}\right)}{\left(n \frac{w_s}{2R_s}\right)} \cdot \sin\left(\frac{n\tau_c}{2}\right) \left(\frac{r}{R_s}\right)^n \cdot \frac{\left(1 + \left(\frac{R_r}{r}\right)^{2n}\right)}{1 - \left(\frac{R_r}{R_s}\right)^{2n}} \cdot \cos(n(\theta - \tau_{coil}/2)) \quad (10.13)$$

For external rotors, a similar formula could be obtained with $\frac{R_s}{r}$ instead of $\frac{r}{R_s}$, $\frac{r}{R_r}$ instead of $\frac{R_r}{r}$ and $\frac{R_s}{R_r}$ instead of $\frac{R_r}{R_s}$.

Ultimately, both PM and mmf airgap flux density expressions in Equation 10.9 and Equation 10.10 have to be multiplied by the P.U. airgap permeance function $\lambda_{p.u.}(\theta)$ of Equation 10.7 with the contributions of all coils for each phase added up (Equation 10.3 and Equation 10.4).

Denoting by E_a , E_b , and E_c the total emf per phases a , b , and c , the electromagnetic torque T_e is, simply,

$$T_e = \frac{(E_a i_a + E_b i_b + E_c i_c)}{\omega_1} \cdot p_1; \quad \omega_1 = \Omega_r \cdot p_1 \approx \frac{U}{R_m} \cdot p_1 \quad (10.14)$$

The instantaneous electromagnetic torque includes the interaction torque between the PM flux and stator current and the so-called reluctance (no PM) torque.

The torque at zero current (cogging torque) is not, however, included in Equation 10.14. The cogging torque is, in fact, produced by the tangential forces on the slot walls:

$$T_{cog}(\theta) \equiv \frac{\pi l_{stack} R_s}{2\mu_0 N_s} \cdot \sum_{m=1}^{N_s} \left[B_{PM}^2 \left(\frac{2\pi m}{N_s} + \theta_1 \right) \cdot (R_m + g_\alpha) ssg \right] \quad (10.15)$$

$g_\alpha = 0$ and $ssg = 0$ outside the slot opening; $g_\alpha = w_1 + g$ and $ssg = 1$ on the left side of the slot opening; and $g_\alpha = w_2 + g$, and $ssg = -1$ on the right side of the slot opening.

Typical results that compare PM, mmf airgap flux densities, cogging, and total torque obtained from the analytical model and from 2D FEM are shown in Figure 10.12a through Figure 10.12d [12]. The 2D analytical model looks adequate to describe airgap flux density pulsations due to slot openings both for the PM and armature contributions. Also, it correctly portrays the cogging torque and the total torque variation with rotor position (or with time). Any shape of current waveform can be dealt with through its time harmonics.

The model, however, ignores magnetic saturation and cannot directly handle inset or interior PM rotor configurations.

As already alluded to, a similar quasi-3D analytical model was developed for disk-shaped (axial flux) PM rotor configurations [13, 14].

The 2D analytical model for the axial airgap machine is developed in the axial circumferential coordinates (z , $\theta - \theta_1$), while the correction for the radial direction is handled [13] as follows (Figure 10.13a and Figure 10.13b)

$$g(r) = \frac{1}{\pi} [\tan((r - R_{ai})/a) - \tan^{-1}((r - R_{ae})/a)] \quad (10.16)$$

with R_{ai} and R_{ae} the inner and outer PM radii, and

$$a = \alpha(R_{ae} - R_{ai}) / \tan(\beta\pi/2) \quad (10.17)$$

with α and β as correction coefficients. By making the computations for 15 to 20 values of the radius and averaging them, while taking into account that the pole pitch τ and the PM span τ_m vary with radius r , similar results may be obtained.

The emf and cogging torque waveforms obtained through such a model [14] are compared with 3D FEM results (Figure 10.14a and Figure 10.14b). There is some 11% difference between analytical and 3D

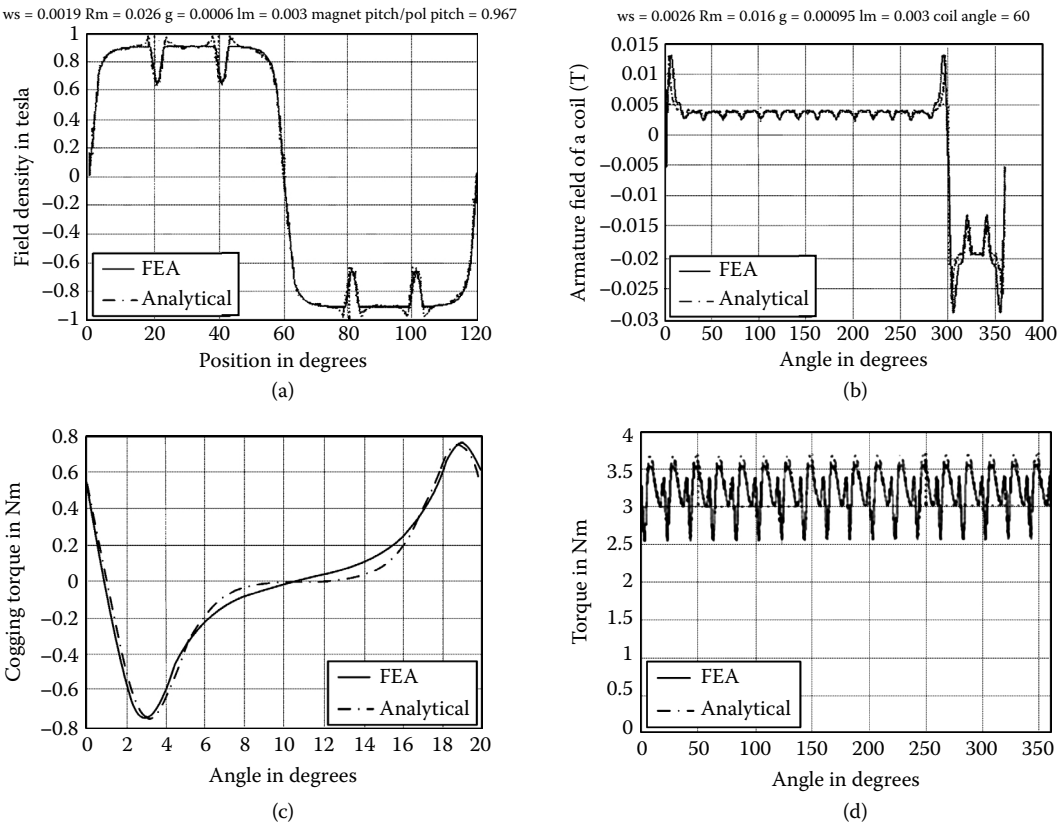


FIGURE 10.12 Analytical vs. finite element method (FEM) airgap flux density distributions and torques for sinusoidal current: (a) permanent magnet (PM) flux density, (b) armature flux density, (c) cogging torque, and (d) total torque.

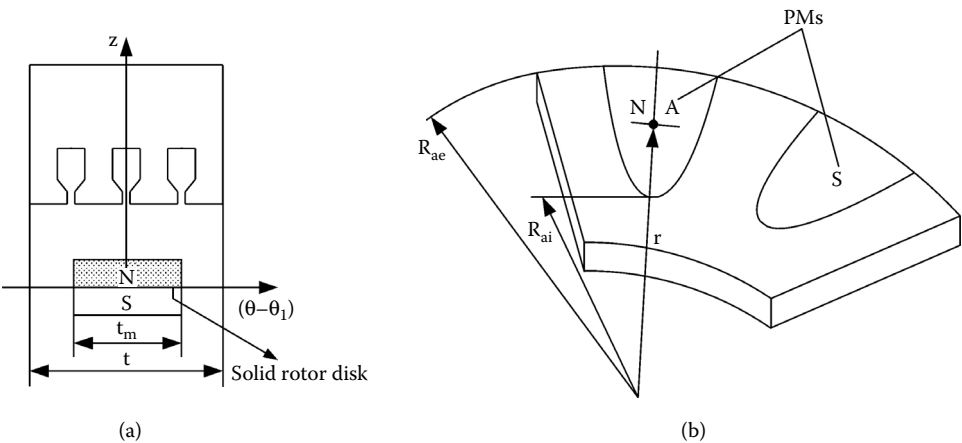


FIGURE 10.13 Disk-rotor permanent magnet synchronous generator (PMSG) model: (a) two-dimensional geometry (z , $\theta - \theta_1$) and (b) radial (third) dimension r .

FEM results, probably due to the fringing flux through stator slot necks which is neglected. Unfortunately, this fringing (leakage) PM flux may be calculated only through FEM. FEM-based corrections for fringing are possible in an iterative way or through applying a fudge factor from experiments. The same assertion should hold for the cylindrical rotor analytical model. The situation is delicate, especially with small slot openings.

Note that while a complete 2D (3D) analytical field model was developed for both cylindrical and disk-shaped rotors, its Fourier series nature makes it useful for design optimization (Figure 10.14a and Figure 10.14b). It is still a bit too complicated for preliminary design, when a more simplified circuit model with analytically defined parameters is required. Finally, the 2D (3D) FEM may be used for model design optimization.

But before dealing with the circuit model, let us explore first cogging torque details and core loss modeling, as they make use of FEM, also, in advanced representations.

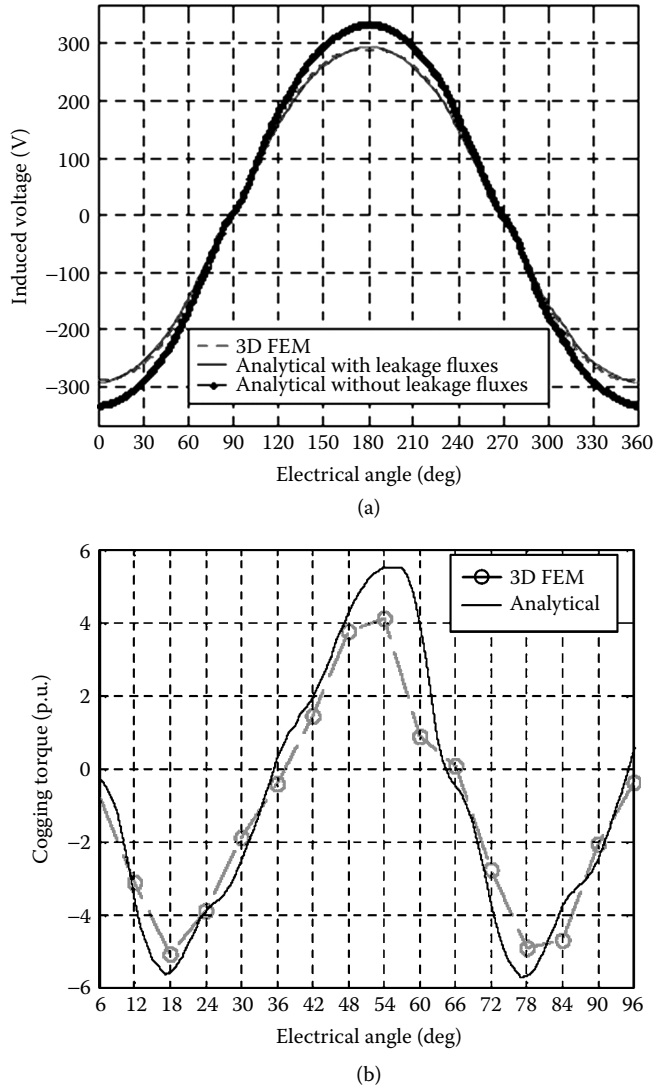


FIGURE 10.14 Disk-shaped rotor permanent magnet synchronous generator (PMSG) by three-dimensional analytical vs. three-dimensional finite element method models: (a) electromagnetic field (emf) waveforms and (b) cogging torque.

10.4 Stator Core Loss Modeling

There are two ways to approach the complex problem of stator core losses in PMSGs. The first approach starts with a simplified flux distribution and a simplified core loss formula and eventually advances to using FEM for refinements.

In the second approach, FEM distribution of flux density evolution with time is considered in each core volume element, which gives stator current/time waveforms. Then, the hysteresis and eddy current losses are added up to obtain the total core losses by again using some simplified core loss formulas.

Recent research efforts showed that not only is the flux density not sinusoidal in time at a point in a stator core, but also its vector direction changes in time at that point, more in the yoke and less in the teeth. This way, the traveling component of flux density time variation was proven to be important in computing the core loss. The data from AC magnetization (from Epstein probe) have to be corrected with the eddy current losses due to rotation. The presence of slot openings and the distribution of windings in slots lead to time harmonics in the stator core flux density, even for distributed windings and sinusoidal currents at constant speed. Also, the on-load core losses tend to differ from no-load core losses; thus, the simplified core loss standard formulas [3, 5] may be used only in very preliminary design stages. The presence of slot openings and the distribution of windings in slots lead to time harmonics in the stator core flux density, even for distributed windings and sinusoidal currents at constant speed. Finally, the importance of “rotation” on core loss tends to increase for concentrated windings in comparison with distributed windings.

All of these aspects tend to suggest that a realistic approach should start first with FEM core loss estimates, and then their curve fitting to even more simplified formula may be operated, for design optimization tasks.

10.4.1 FEM-Derived Core Loss Formulas

It has been known for a long time that the combined alternative and traveling fields produce more core losses than alternating fields of the same frequency and amplitude. The quantization of this phenomenon has not been, in general, taken into consideration.

FEM has brought in a possibility to account for rotational field losses in core losses of brushless machines (Figure 10.15a through Figure 10.15e) [15]. In different points, such as a, b, c, d, e, the flux density has traveling field components of various values, depending on their locations in the stator teeth and yoke. Further on, at some points, the situation is different for given points for concentrated windings (Figure 10.16a and Figure 10.16b) [15].

It is evident that the short/long axis ratio c of the elliptical variation of B_t/B_o for each harmonic is different for distributed and concentrated windings (Figure 10.15a through Figure 10.15d, and Figure 10.16a and Figure 10.16b).

For the distributed winding, the traveling field components ($c > 0.3$) are notable at the teeth bottom and in part of the yoke, while for concentrated winding, they also show up near the teeth top and throughout the yoke. Larger core losses are expected for concentrated windings. A general FEM-derived formula for total stator core losses is of the following form [15]:

$$P_{Fe-total} = \sum_{j=1}^n \left\{ \sum_{i=1}^m g_i \left[\left(\epsilon_h \left(\frac{jf}{100} \right) B_{mji}^\alpha + \epsilon_{eddy} \left(\frac{jf}{100} \right)^2 B_{mji}^\alpha \right) \cdot (1 + \gamma c_{ij}) \right] \right\} \quad (10.18)$$

where

j is the order of the flux density time harmonic

i is the finite element number

g_i is the mass of i finite element

f is the frequency

γ is a correcting constant

c is the short-to-long axis ratio of field vector hodograph in element i for the harmonic j

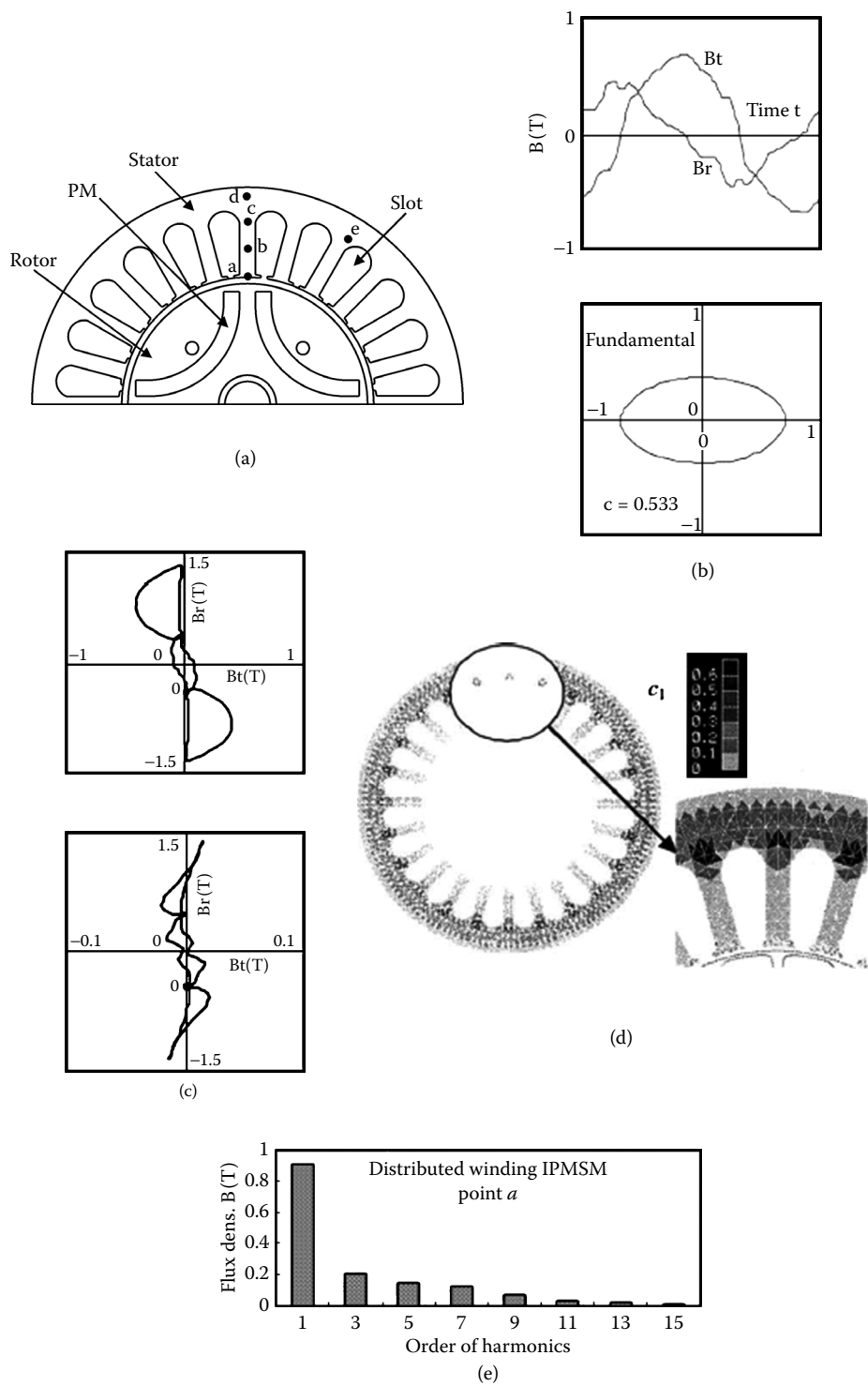


FIGURE 10.15 Distributed winding interior permanent magnet synchronous generator (IPM-SG) core flux density: (a) the points a, b, c, d, and e, (b) flux density vector hodograph in point a, (c) elliptical variation for each harmonic, (d) short to long axis c coefficient, and (e) harmonics spectrum of flux density.

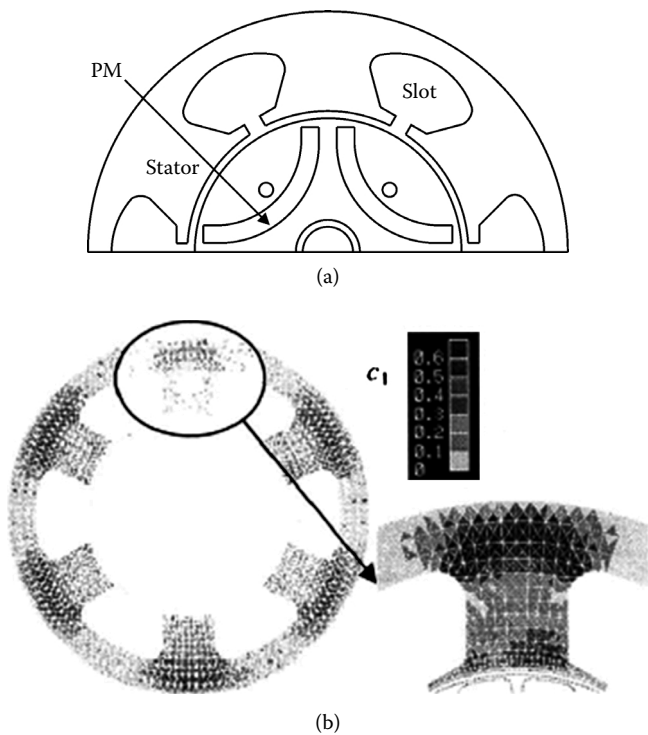


FIGURE 10.16 Concentrated winding interior permanent magnet synchronous generator (IPM-SG): (a) cross-section and (b) distribution of short to long axis rate c .

In case stator current time harmonics are considered, the same formula is applied for each current harmonic frequency.

Comparison between experimental losses and calculated ones by conventional (analytical) models and Equation 10.18 — for the distributed and concentrated windings — are shown in Figure 10.17 and Figure 10.18 [15].

The increase in precision by considering traveling field components is evident.

Also, the contribution of harmonics in core losses is shown in Figure 10.19 and Figure 10.20 for sinusoidal current.

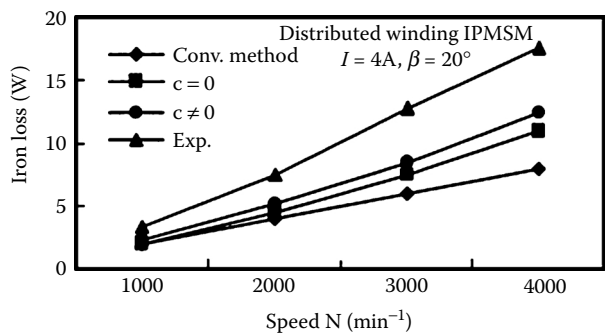


FIGURE 10.17 Core loss evaluation for distributed winding with load.

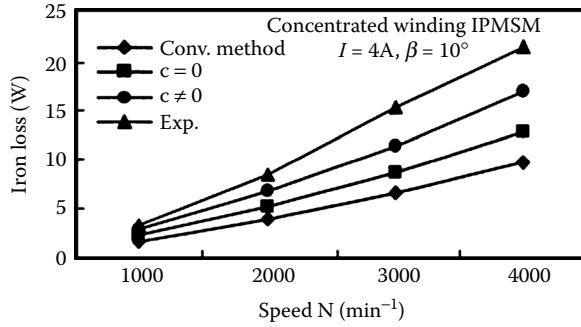


FIGURE 10.18 Core loss evaluation for concentrated windings with load.

As expected, the third flux density harmonic is responsible for about 25% of core loss for the concentrated winding. Extending the use of concentrated windings at high speed should be done with care, as core losses tend to be larger than those for distributed windings.

Alternatively, by adding the core loss of radial and tangential flux density variations in each finite element for N samples in a time period and over all finite mass elements and separately for eddy currents and hysteresis losses, another total core loss formula may be obtained [16]:

$$P_{Fe-total} = \sum_{i=1}^m \frac{g_i}{N} \sum_{k=1}^N k_e \left[\left(\frac{B_r^{k+1} - B_r^k}{\Delta t} \right)^2 + \left(\frac{B_\theta^{k+1} - B_\theta^k}{\Delta t} \right)^2 \right] + \sum_{i=1}^m \frac{g_i}{T} \cdot \frac{K_h}{2} \cdot \left[\sum_{j=1}^{N_{pr}^i} (B_{mr}^{ij})^2 + \sum_{j=1}^{N_{p\theta}^i} (B_{m\theta}^{ij})^2 \right] \quad (10.19)$$

where

Δt is the sampling time (there are N sampling times in one period T)

B_r^k, B_θ^k are the radial and tangential instantaneous flux density components in element i

$B_{mr}^{ij}, B_{m\theta}^{ij}$ are the amplitudes of radial and axial flux densities in element i for the hysteresis cycle j

k_e, k_h are experimental constants from the Epstein frame obtained from the slope and intercept variation of core losses/frequency vs. frequency straight line

Equation 10.19 may also be applied for the computation of rotor core losses.

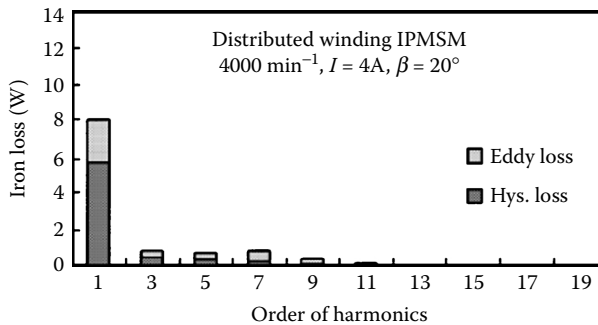


FIGURE 10.19 Flux density harmonics contributions to core losses (distributed winding).

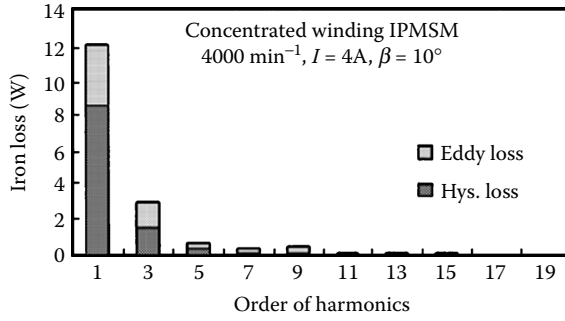


FIGURE 10.20 Core loss harmonics breakdown for concentrated winding.

The FEM-derived formulas account for actual flux density evolution in all parts of the magnetic circuit, either for no-load or for on-load conditions. Also, the case of rectangular or sinusoidal current waveforms may be treated with the same method.

Finally, IPM or surface PM rotors can be accommodated by simplified formulas.

10.4.2 Simplified Analytical Core Loss Formulas

Analytical core loss formulas for stator core loss in PMSGs were derived first for distributed windings and sinusoidal currents. It was implicit that the flux density varies sinusoidally in time in both stator teeth and yoke, and also, that their value is the same in all points of the core. Both these approximations are coarse, as seen in the previous paragraph. The no-load core losses were evaluated first. A simple formula that even ignores the hysteresis losses is as follows:

$$P_{Fe} / kg \approx 2.8 \cdot k_e \cdot \omega^2 \cdot \hat{B}_g^2 \quad (10.20)$$

$\hat{B}_g^2$ is the peak flux density in the airgap. As this formula cannot discriminate between lightly saturated or heavily saturated teeth and yoke designs, it is somewhat surprising that it is still mentioned in the literature today [5]. But as long as the core losses are notably smaller than copper losses, notable errors in the former do not produce design disasters. This is not so in optimized designs, where winding and nonwinding losses tend to be of about the same value, or in high-speed PMSGs.

For $q = 1$ distributed windings and rectangular current, the PM airgap flux density distribution produces different, linear, flux density variations in the stator teeth and in the yoke (Figure 10.21) at no load [3]. The total eddy current loss formula at no load for such cases is of the following form:

$$P_{eddy} / kg = \frac{4R}{\pi p_1} k_e \omega^2 \left(\frac{2(w_t + w_s)^2}{w_t^3} + \frac{\alpha R}{p \cdot hys^2} \right) \hat{B}_g^2 \quad (10.21)$$

where

R is the rotor radius

w_t is the tooth width

w_s is the slot width

hys is the stator yoke radial thickness

p_1 is the pole pairs

ω is the electrical fundamental frequency

$\hat{B}_g^2$ is the peak flux density in the airgap

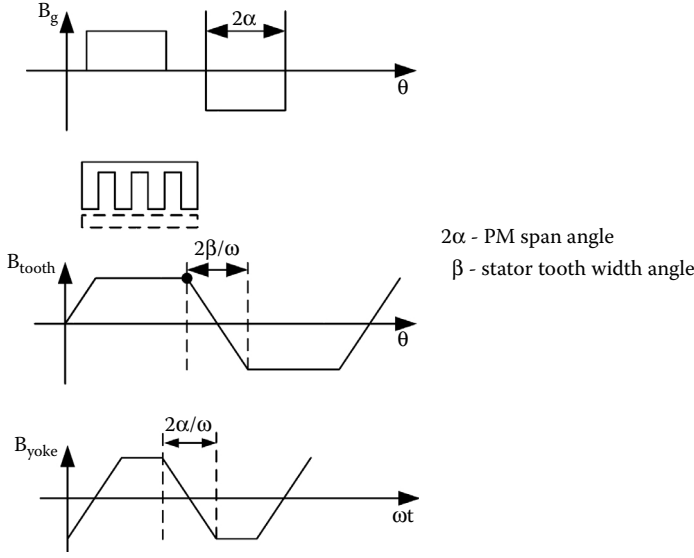


FIGURE 10.21 Ideal teeth and yoke flux densities variation with rotor position for rectangular airgap flux density.

A rough approximation of Equation 10.21 is as follows [16]:

$$P_{eddy} \approx 6 \cdot k_e \cdot \left(\omega^2 \cdot \hat{B}_g^2 \right) \quad (10.22)$$

Equation 10.20 and Equation 10.22 show that the rectangular airgap flux density in the airgap is characterized by notably larger core loss than a sinusoidal one. Consequently, when core loss becomes important, sinusoidal airgap flux density distribution should be targeted. In the above analytical expressions, the armature reaction field was not considered. In heavily loaded and, especially in IPM rotors, such an approximation is hardly practical.

Also, it is to be noted that eddy current core losses depend on the flux density waveform (or its variation velocity). The hysteresis core losses depend essentially only on the peak values of flux density. This is how simplified analytical formulas for eddy current and hysteresis core loss for rectangular flux density variation in time and IPM-SG on load were derived [17]:

$$P_{eddy} = P_{etooth} + P_{eyoke} \quad (10.23)$$

$$P_{eteeth} / kg \approx \frac{k_e}{T} \int_0^T \left(\frac{dB_t}{dt} \right)^2 dt = k_e \frac{\omega^2}{\pi} \left[\frac{2}{p_1 \cdot \beta} \left((B_{gm} + B_{dm} - \hat{B}_d \cos \alpha_t)^2 + \hat{B}_q^2 \sin^2 \alpha_t \right) + \hat{B}_d^2 \left(\frac{\alpha_t - \sin 2\alpha_t}{2} \right) + \hat{B}_q^2 \left(\frac{\alpha_t + \sin 2\alpha_t}{2} \right) \right] \quad (10.24)$$

$$P_{eyoke} / kg \approx \frac{k_e}{\pi} \left(\frac{R_s \omega}{p_1 \cdot hys} \right)^2 \left[2p_1 \alpha (B_{gm} + B_{dm})^2 + \hat{B}_d^2 \left(p_1 \alpha + \frac{\sin 2p_1 \alpha}{2} \right) + \hat{B}_q^2 \left(\frac{2p_1 \alpha - \sin 2p_1 \alpha + p_1 \delta + \sin p_1 \alpha}{2} \right) - 4(B_{gm} + B_{dm}) \hat{B}_d \sin p_1 \alpha \right] \quad (10.25)$$

where

- B_{gm} is the PM airgap flux density
- $\hat{B}_{dm}$ is the peak airgap flux density at the rotor surface
- $\hat{B}_d$ is the peak airgap flux density due to d axis current
- $\hat{B}_q$ is the peak airgap flux density due to q axis current
- α is the pole magnet mechanical angle
- $\alpha_t = p_1(\alpha - \beta/2)$
- β is the slot pitch mechanical angle
- R_s is the stator interior radius

Finally, the hysteresis losses are expressed by the usual formula:

$$P_{h_{t,y}}/kg = k_h \cdot \omega_1 \cdot \hat{B}_{t,y}^2 \quad (10.26)$$

Though expressions such as Equation 10.24 through Equation 10.26 tend to account for the load (armature reaction) influence on core losses, they still cannot discriminate directly between designs with various teeth thickness (geometry) or local magnetic saturation effects. However, as shown in Figure 10.22 [17], the load has a strong influence on stator core losses. In terms of core losses, FEM verifications at the design stage are highly recommended for critical operation modes. Experiments should follow whenever possible.

Note that it was suggested that core losses be represented in the circuit model of PMSG by an equivalent resistance R_c of the following form:

$$R_c = \frac{1}{\frac{1}{R_{eddy}} + \frac{1}{\omega_1 r_{hys}}} = \frac{\frac{2}{3} E_{total}^2}{P_{Fe_{total}}} \quad (10.27)$$

E_{total} is the total emf in the d - q model of PMSG. The term ω_1 accounts for the hysteresis loss linear variation with frequency. E_{total} is also proportional to frequency. Eddy current core losses are proportional to frequency squared. Such formulas should also be used with care, as, in fact, R_{eddy} and r_{hys} vary in turn with load conditions, and so forth.

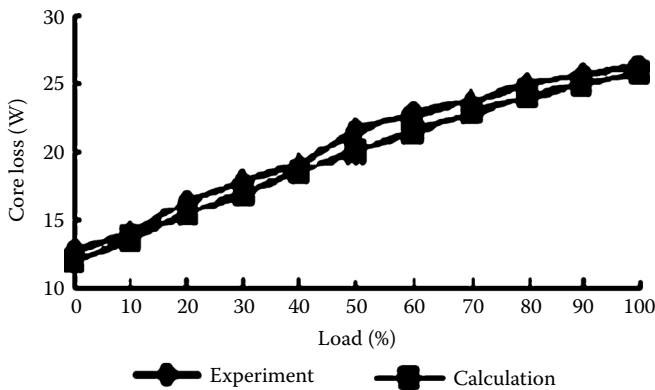


FIGURE 10.22 Core loss vs. load at 2500 rpm for a four-pole three-phase 600 W interior permanent magnet (IPM) synchronous machine.

For transients, it seems better to neglect stator core losses, while for steady state, it is more practical to add them, after their FEM computation or measurement, to estimate efficiency or temperatures in the machine.

Note that rotor losses were mentioned only with respect to rotor core and in relation to FEM. In principle, eddy currents occur in the PMs due to armature field space harmonics. In high-speed machines, they are notable [26].

10.5 The Circuit Model

10.5.1 The Phase Coordinate Model

In essence, the circuit model of a PMSG starts with the phase voltage equations in stator coordinates:

$$\begin{aligned} i_a R_s - V_a &= -\frac{d\Psi_a}{dt} \\ i_b R_s - V_b &= -\frac{d\Psi_b}{dt} \\ i_c R_s - V_c &= -\frac{d\Psi_c}{dt} \end{aligned} \quad (10.28)$$

$$\begin{bmatrix} \Psi_a \\ \Psi_b \\ \Psi_c \end{bmatrix} = \begin{bmatrix} L_{sl} + L_{aa}(\theta_{er}) & L_{ab}(\theta_{er}) & L_{ca}(\theta_{er}) \\ L_{ab}(\theta_{er}) & L_{sl} + L_{bb}(\theta_{er}) & L_{bc}(\theta_{er}) \\ L_{ca}(\theta_{er}) & L_{bc}(\theta_{er}) & L_{sl} + L_{cc}(\theta_{er}) \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + \begin{bmatrix} \Psi_{PMa}(\theta_{er}) \\ \Psi_{PMb}(\theta_{er}) \\ \Psi_{PMc}(\theta_{er}) \end{bmatrix} \quad (10.29)$$

where θ_{er} is the rotor PM axis angle to stator phase a axis/electrical angle.

The self-inductance and mutual inductance of the stator depend sinusoidally on θ_{er} only for IPM rotors, in distributed winding IPM rotor machines. For surface PM pole rotors, stator inductances are independent of θ_{er} . However, the presence of slot openings introduces an additional dependence of stator inductances on $N_s \theta_{er}$ (N_s is the number of stator slots) for IPM rotors.

For concentrated windings, the stator self-inductance and mutual inductance perform in a similar way, with respect to rotor pole configurations, but their values tend to be larger than for distributed windings and equivalent machine geometries. However, the end-turn leakage inductances are notably smaller for concentrated windings.

The PM flux linkages in the stator phases $\Psi_{PMa}(\theta_{er})$, $\Psi_{PMb}(\theta_{er})$, and $\Psi_{PMc}(\theta_{er})$ are either sinusoidal or they also contain space harmonics.

The trapezoidal distribution may be considered this way:

$$\Psi_{PMa}(\theta_{er}) = \Psi_{PM1}(\theta_{er}) + \Psi_{PM2}(2\theta_{er} - \gamma_2) + \Psi_{PM3}(3\theta_{er} - \gamma_3) + \dots \quad (10.30)$$

The even harmonics occur only with PM pole shifting, adopted to reduce cogging torque. Also, space subharmonics may occur. They have to be avoided for practical designs.

For surface PM pole rotors, the general expressions of E_{aa} , E_{ab} , E_{ac} , and Ψ_{PMa} developed earlier may be used to calculate L_{aah} , L_{abh} , L_{cah} , Ψ_{PMa} , ($i_b = i_c = 0$):

$$L_{aah} = \frac{E_{aa}}{\omega_1 i_a}; L_{abh} = \frac{E_{ab}}{\omega_1 i_a}; L_{cah} = \frac{E_{ac}}{\omega_1 i_a}; \Psi_{PMa} = \frac{E_{PMa}}{\omega_1} \quad (10.31)$$

The analytical method in Section 10.2 may be applied to estimate all parameters in Equation 10.29 and Equation 10.30 for distributed and concentrated windings for surface PM pole rotor configurations.

FEM field distributions may produce even better values for inductances and $\Psi_{PMa,b,c}(\theta_{er})$ expressions, and they are also recommended for IPM rotors.

The instantaneous interaction torque expression, in the absence of magnetic saturation, is as follows:

$$T_e = -\frac{\partial W_m}{\partial \theta_{er}} \quad (10.32)$$

$$W_m = \frac{1}{2} L_{aah} i_a^2 + \frac{1}{2} L_{bbh} i_b^2 + \frac{1}{2} L_{cch} i_c^2 + L_{abh} i_a i_b + L_{bch} i_b i_c + L_{cah} i_c i_a \\ + \Psi_{PMa}(\theta_{er}) i_a + \Psi_{PMb}(\theta_{er}) i_b + \Psi_{PMc}(\theta_{er}) i_c \quad (10.33)$$

Though it is feasible to use the phase coordinate model, especially for complex digital simulations in the presence of static power converters, and, even for trapezoidal emf distributions, the d - q model is preferred when simpler models for control design are required. The PM flux linkage harmonics are introduced to simulate its departure from a sinusoidal distribution.

The d - q model with emf (PM emf) space harmonics operates as a general tool for design and control, for both sinusoidal and trapezoidal emf and distributed or concentrated windings, either for surface pole PM or for interior PM rotor pole configurations. For concentrated windings and IPM pole rotors, sinusoidal PM flux linkage distributions are targeted, and again, the d - q model is feasible.

10.5.2 The d - q Model of PMSG

The d - q model is based on the assumption that the stator self-inductance and mutual inductance are either constant or vary sinusoidally with the rotor position ($2\theta_{er}$). In general, the PM flux linkages $\Psi_{PMa,b,c}(\theta_{er})$ in the stator phases also vary sinusoidally, but with θ_{er} . Eventual harmonics in $\Psi_{PMa,b,c}(\theta_{er})$ may be treated in the d - q model also, and they are expected to create time pulsations in the torque with sinusoidal currents at speed and frequency $\omega_1 = \omega_r$:

$$|L_{abc}(\theta_{er})| = \begin{vmatrix} L_{sl} + L_0 + L_2 \cos 2\theta_{er} & M_0 + L_2 \cos\left(2\theta_{er} + \frac{2\pi}{3}\right) & M_0 + L_2 \cos\left(2\theta_{er} - \frac{2\pi}{3}\right) \\ M_0 + L_2 \cos\left(2\theta_{er} + \frac{2\pi}{3}\right) & L_{sl} + L_0 + L_2 \cos\left(2\theta_{er} - \frac{2\pi}{3}\right) & M_0 + L_2 \cos 2\theta_{er} \\ M_0 + L_2 \cos\left(2\theta_{er} - \frac{2\pi}{3}\right) & M_0 + L_2 \cos 2\theta_{er} & L_{sl} + L_0 + L_2 \cos\left(2\theta_{er} + \frac{2\pi}{3}\right) \end{vmatrix} \\ M = -\frac{L_0}{2} \text{ for distributed windings} \quad (10.34)$$

For PM machines, $L_2 < 0$, as they have the PMs placed along axis d and exhibit “inverse saliency,” in contrast to standard synchronous machine excitation:

$$\begin{vmatrix} \Psi_{PMa}(\theta_{er}) \\ \Psi_{PMb}(\theta_{er}) \\ \Psi_{PMc}(\theta_{er}) \end{vmatrix} = \begin{vmatrix} \Psi_{PM1} \cos \theta_{er} + \dots \\ \Psi_{PM1} \cos\left(\theta_{er} - \frac{2\pi}{3}\right) + \dots \\ \Psi_{PM1} \cos\left(\theta_{er} + \frac{2\pi}{3}\right) + \dots \end{vmatrix} \quad (10.35)$$

The matrix form of the phase coordinates model is as follows:

$$\left| i_{a,b,c} \right| R_s - \left| V_{a,b,c} \right| = - \frac{d \left| \Psi_{a,b,c} \right|}{dt} \quad (10.36)$$

$$\Psi_{a,b,c} = \left| L_{a,b,c}(\theta_{er}) \right| i_{a,b,c} + \Psi_{PMa,b,c}(\theta_{er}) \quad (10.37)$$

The Park transformation $P(\theta_{er})$ is used to derive the d - q model:

$$\left| P(\theta_{er}) \right| = \frac{2}{3} \begin{vmatrix} \cos(-\theta_{er}) & \cos\left(-\theta_{er} + \frac{2\pi}{3}\right) & \cos\left(-\theta_{er} - \frac{2\pi}{3}\right) \\ \sin(-\theta_{er}) & \sin\left(-\theta_{er} + \frac{2\pi}{3}\right) & \sin\left(-\theta_{er} - \frac{2\pi}{3}\right) \\ \frac{1}{2} & \frac{1}{2} & \frac{1}{2} \end{vmatrix} \quad (10.38)$$

The Park transformation from stator to rotor coordinates is, in Equation 10.38, valid for the trigonometric motion direction and axis q in front of axis d by 90° (electrical degrees) (Figure 10.23):

$$\begin{vmatrix} i_d \\ i_q \\ i_0 \end{vmatrix} = \left| P(\theta_{er}) \right| \begin{vmatrix} i_a \\ i_b \\ i_c \end{vmatrix} \quad (10.39)$$

The same transformation is valid for Ψ_{dq0}, V_{dq0} .

Finally, for sinusoidal $\Psi_{PMa,b,c}(\theta_{er})$ distributions,

$$\begin{aligned} i_d R_s - V_d &= -L_d \frac{di_d}{dt} + \omega_r L_q i_q \\ i_q R_s - V_q &= -L_q \frac{di_q}{dt} - \omega_r (L_d i_d + \Psi_{PM1}) \end{aligned} \quad (10.40)$$

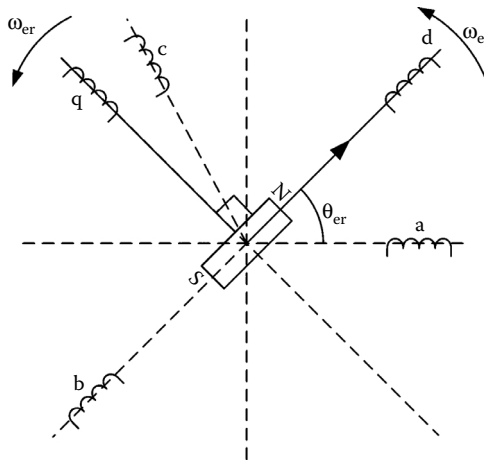


FIGURE 10.23 Three-phase to d - q transformation.

with

$$\begin{aligned}\bar{\Psi}_s &= \Psi_d + j\Psi_q & \Psi_d &= \Psi_{PM1} + L_d i_d & \Psi_q &= L_q i_q \\ \bar{V}_s &= V_d + jV_q & \bar{i}_s &= i_d + j i_q\end{aligned}\quad (10.41)$$

The so-called space-vector (or complex variable) model of a PMSG is obtained:

$$\bar{i}_s R_s - \bar{V}_s = -\frac{d\bar{\Psi}_s}{dt} - j\omega_r \bar{\Psi}_s \quad (10.42)$$

The torque is obtained from power balance in Equation 10.37:

$$T_e = p_1 \frac{P_e}{\omega_r} = \frac{3}{2} p_1 \operatorname{Re}(j \bar{\Psi}_s \bar{i}_s^*) = \frac{3}{2} p_1 (\Psi_d i_q - \Psi_q i_d) = \frac{3}{2} p_1 (\Psi_{PM1} + (L_d - L_q) i_d) i_q \quad (10.43)$$

Also,

$$L_d = L_s + \frac{3}{2}(L_0 - |L_2|) \quad L_q = L_s + \frac{3}{2}(L_0 + |L_2|) \quad (10.44)$$

The winding losses P_{copper} are as follows:

$$P_{copper} = \frac{3}{2} R_s (i_d^2 + i_q^2) \quad (10.45)$$

The expressions in Equation 10.40 lead to the d - q circuit model in Figure 10.24, where the association of signs corresponds to motoring. For generating, i_q changes sign (is negative) and so does the torque. The corresponding general vector diagrams for motoring and generating are shown in Figure 10.25a and Figure 10.25b.

Under steady state, $s = 0$ (rotor coordinates), and the circuit model and vector diagram further simplify. For steady state and sinusoidal phase voltages,

$$V_{abc} = V_1 \sqrt{2} \cos\left(\omega_r t - (i-1) \frac{2\pi}{3}\right) \quad (10.46)$$

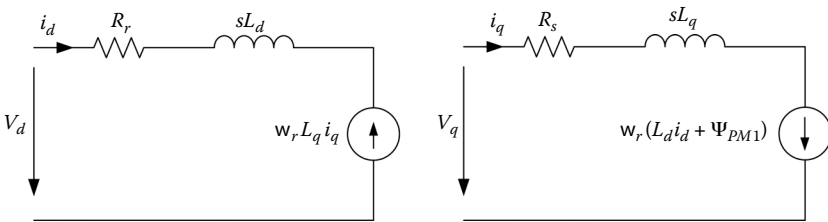


FIGURE 10.24 d - q equivalent circuits of permanent magnet synchronous generator induction machine (PMSGIM).

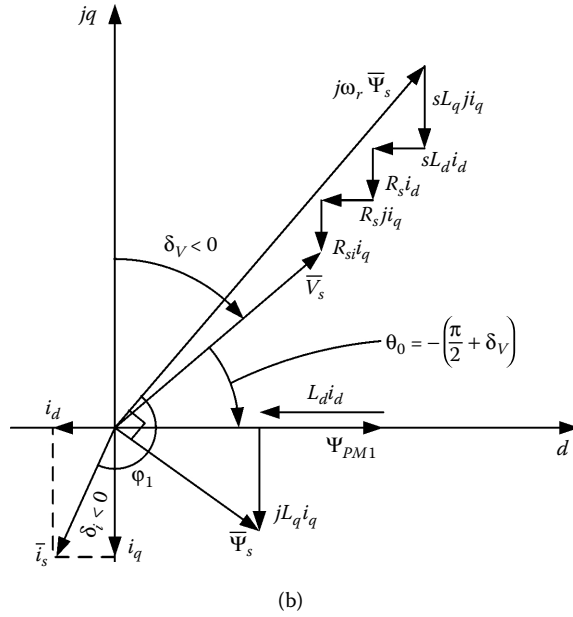
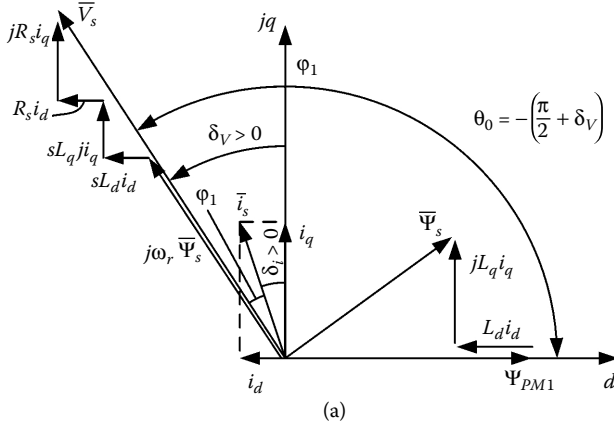


FIGURE 10.25 General vector diagrams of permanent magnet synchronous generator induction machine (PMSG): (a) motoring and (b) generating.

According to the Park transformation, where $\theta_{er} = \omega_r t + \theta_0$ ($\omega_r = \text{constant}$) and Figure 10.25b,

$$\begin{aligned} V_d &= V_1 \sqrt{2} \cos \theta_0; \quad \theta_0 = -\left(\frac{\pi}{2} + \delta_V\right) \\ V_q &= -V_1 \sqrt{2} \sin \theta_0 \end{aligned} \quad (10.47)$$

$\delta_V > 0$ for motoring and $\delta_V < 0$ for generating, or $V_d < 0, V_q > 0$ for motoring, and $V_d > 0, V_q < 0$ for generating. Also, $i_d < 0$ and $i_q > 0$ for motoring, and $i_q < 0$ for generating. The power factor angle $0 < \phi_1 < 90^\circ$ for motoring, and $\phi_1 > 90^\circ$ for generating. To produce reactive power in generating, $\phi_1 > 180^\circ$.

Example 10.1

Consider a surface PM pole rotor PMSG with $R_s = 1\Omega$, $L_s = L_d = L_q = 0.05$ H, $\Psi_{PM1} = 0.5$ Wb, and $p_1 = 2$ pole pairs that have to deliver power into a three-phase resistance of $R_L = 10\Omega$ /phase at the speed $n_1 = 3000$ rpm.

Calculate the phase voltage, current, and power delivered to the load resistance. Also, the voltage and current power angles δ_v , δ_i ; justify the results. Change the load resistance to obtain more output power.

Solution

The symmetric three-phase load may be transformed into a d - q load, maintaining $R_L = 10\Omega$.

Thus, the generator turned Equation 10.35, for steady state, is as follows:

$$i_d(R_s + R_L) = \omega_r L_s i_q \quad (A)$$

$$i_q(R_s + R_L) = -\omega_r (\Psi_{PM1} + L_d i_d) \quad (A)$$

$$\omega_r = 2\pi p_1 n_1 = 2\pi \cdot 2 \cdot \frac{3000}{60} = 200\pi$$

Equation (A) allows us to calculate i_d , i_q :

$$i_d(1+10) = +200\pi \cdot 0.05 \cdot i_q$$

$$i_q(1+10) = -200\pi \cdot (0.5 + 0.05 \cdot i_d)$$

$$i_d = -8.9 \text{ A} \quad i_q = -3.12 \text{ A}$$

As the d - q voltage equation was written for motoring,

$$V_d = -R_L i_d = -10 \times (-8.9) = 89 \text{ V}$$

$$V_q = -R_L i_q = -10 \times (-3.12) = 31.2 \text{ V}$$

The phase voltage (RMS) V_1 is as follows:

$$V_1 = \frac{1}{\sqrt{2}} \sqrt{V_d^2 + V_q^2} = \frac{1}{\sqrt{2}} \sqrt{89^2 + 31.2^2} = 66.88 \text{ V}$$

Also,

$$I_1 = \frac{1}{\sqrt{2}} \sqrt{I_d^2 + I_q^2} = \frac{1}{\sqrt{2}} \sqrt{8.9^2 + 3.12^2} = 6.688 \text{ A}$$

The active power, for resistive load P_{out} , is as follows:

$$P_{out} = \frac{3}{2} \cdot (V_d i_d + V_q i_q) = 3 \cdot V_1 \cdot I_1 = 3 \cdot 66.88 \times 6.688 = 1342 \text{ W}$$

The no-load voltage per phase E_{phase} (RMS) is as follows:

$$E_{phase} (RMS) = \frac{\omega_r \Psi_{PM1}}{\sqrt{2}} = \frac{200\pi \cdot 0.5}{\sqrt{2}} = 222.695 \text{ V}$$

The internal voltage power angle δ_V (Figure 10.25b) is as follows:

$$\delta_V = -\tan^{-1} \frac{V_d}{V_q} = -\tan^{-1} \frac{(-8.9)}{-3.12} = -70.68^\circ$$

The voltage power angle δ_V should be equal to the d - q current power angle δ_i , as the load power factor is unity (resistive load):

$$\delta_i = -\tan^{-1} \frac{i_d}{i_q} = -\tan^{-1} \frac{89}{31.2} = -70.68^\circ$$

The “gigantic” voltage regulation (from 222.695 to 66.88 V) is due to the too large $L_d = L_q$. This implies a machine that is too small for the load resistance in consideration.

This fact is certified by $|i_d| > |i_q|$.

Besides resistive loads, diode rectifier loads are also characterized by almost unity power factor, but in this case, the diode commutation process further reduces the rectified voltage as the load increases.

The vector diagram for the case in point is shown in Figure 10.26a and Figure 10.26b. Changing the resistance load to a larger value $R_L = 30 \, \Omega$, we obtain the following:

$$i_d = \frac{-100\pi}{\left(\frac{31^2}{10\pi} + 10\pi\right)} = -5.06 \, A \quad i_q = -5 \, A$$

Again,

$$I_1 = \frac{1}{\sqrt{2}} \sqrt{I_d^2 + I_q^2} = \frac{1}{\sqrt{2}} \sqrt{(-5.06)^2 + (-5)^2} = 5.03 \, A$$

The phase voltage (RMS) V_1 is

$$V_1 = R_L \cdot I_1 = 30 \times 5.03 = 150.90 \, V$$

The output power is as follows:

$$P_{out} = 3V_1 I_1 = 3 \cdot 150.90 \times 5.03 = 2277 \, W$$

A large load resistance per phase led to more output power, as phase voltage ended up larger.

The voltage power angle δ_V is now

$$\delta_i = \delta_V = -\tan^{-1} \frac{i_d}{i_q} = -\tan^{-1} \frac{(-5.06)}{(-5.00)} = -45.34^\circ$$

This is about the maximum power that can be delivered by the machine to a resistive load, as $I_d \approx I_q$ ($L_d = L_q$).

Now the result is more reasonable, but the voltage regulation is still large. Moreover, as rated voltage power angles range around $\delta_V = 30^\circ - 35^\circ$, the rated voltage regulation runs in the range of 30%, which is reasonable for a synchronous machine, even with surface PM pole rotor.

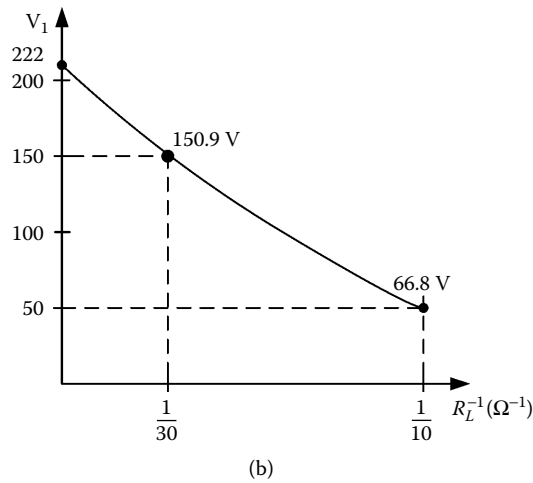
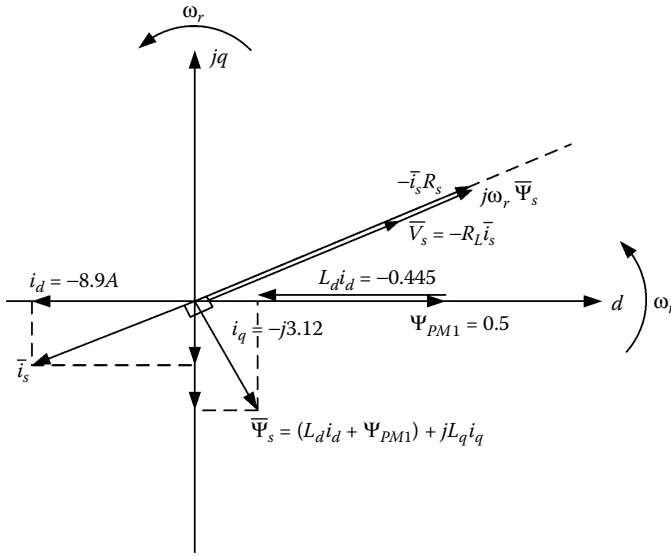


FIGURE 10.26 (a) Unity power factor generator vector diagram for the case in Example 10.1 and (b) voltage vs. R_L^{-1} .

10.6 Circuit Model of PMSG with Shunt Capacitors and AC Load

As visible in Example 10.1, the voltage regulation is notable even for surface PM pole rotor PMSGs with resistive load. IPM-SGs were proposed to make use of inverse saliency ($L_q > L_d$) to bring the terminal voltage back to the no-load voltage for a single (designed) load level. Capacitors in parallel or in series may be connected at PMSG terminals to compensate for increased machine required reactive power as the load level increases.

Parallel capacitors are considered here (Figure 10.27).

To model the PMSG plus shunt capacitors and load, making use of the d - q model, the capacitors and the load have to be transformed into the d - q model.

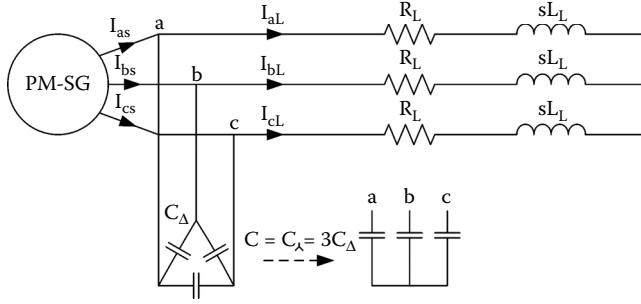


FIGURE 10.27 Permanent magnet synchronous generator (PMSG) with shunt capacitors and alternating current (AC) load.

In phase coordinates, the capacitor and load equations are as follows:

$$\frac{dV_{a,b,c}}{dt} = \frac{(i_{a_s,b_s,c_s} - i_{a_L,b_L,c_L})}{C} \quad (10.48)$$

$$V_{a,b,c} = -(R_L + sL_L)i_{a_L,b_L,c_L} \quad (10.49)$$

The sign (−) written for motoring association of signs in Equation 10.48 reflects the fact that the PMSG d – q model is transformed into d – q rotor coordinates by using the inverse Park transform $\frac{3}{2}[P(\theta_{er})]^T$ (Equation 10.38) to obtain [18]:

$$sV_d = (I_d - I_{dL})\frac{1}{C} + \omega_r V_q \quad (10.50)$$

$$sV_q = (I_q - I_{qL})\frac{1}{C} - \omega_r V_d$$

$$sI_{dL} = -\frac{1}{L_L}V_d - \omega_r I_{qL} - \frac{R_L}{L_L}i_{qL} \quad (10.51)$$

$$sI_{qL} = -\frac{1}{L_L}V_q + \omega_r I_{dL} - \frac{R_L}{L_L}i_{dL}$$

Now we simply add the PMSG d – q model equations derived in the previous paragraph:

$$\begin{aligned} s\Psi_d &= V_d - R_s i_d + \omega_r \Psi_q & \Psi_q &= L_d i_q \\ s\Psi_q &= V_q - R_s i_q - \omega_r \Psi_d & \Psi_d &= L_d i_d + \Psi_{PM1} \end{aligned} \quad (10.52)$$

The currents i_d, i_q may be expressed as functions of Ψ_d and Ψ_q and are thus dummy variables. The actual variables are $I_{dL}, I_{qL}, V_d, V_q, \Psi_d, \Psi_q$. For steady state, $s = 0$ (rotor coordinates). While for surface PM pole rotors, $L_d = L_q \approx \text{const}$, for IPM rotors both $L_d \neq L_q$ and Ψ_{PM1} may vary with magnetic saturation.

In Reference [19], it is assumed that L_d, L_q, Ψ_{PM1} depend on the stator current vector i_s , that is cross-coupling saturation is implicit and based on the assumption of two unique d – q axis curves. As both L_d and L_q decrease with i_s , the voltage regulation drops with increased loads.

A maximum power point on the voltage/power curve is obtained. The shunt capacitors at constant speed move this maximum to higher values for even smaller voltage regulation as in Figure 10.28a and Figure 10.28b[19]. For constant $L_d = L_q$ and Ψ_{PM1} parameters, Equation 10.50 through Equation 10.52 yield straightforward solutions for $V_d, V_q, i_d, i_q, i_{dL}, i_{qL}$ for given C, L_L, R_L , and ω_r .

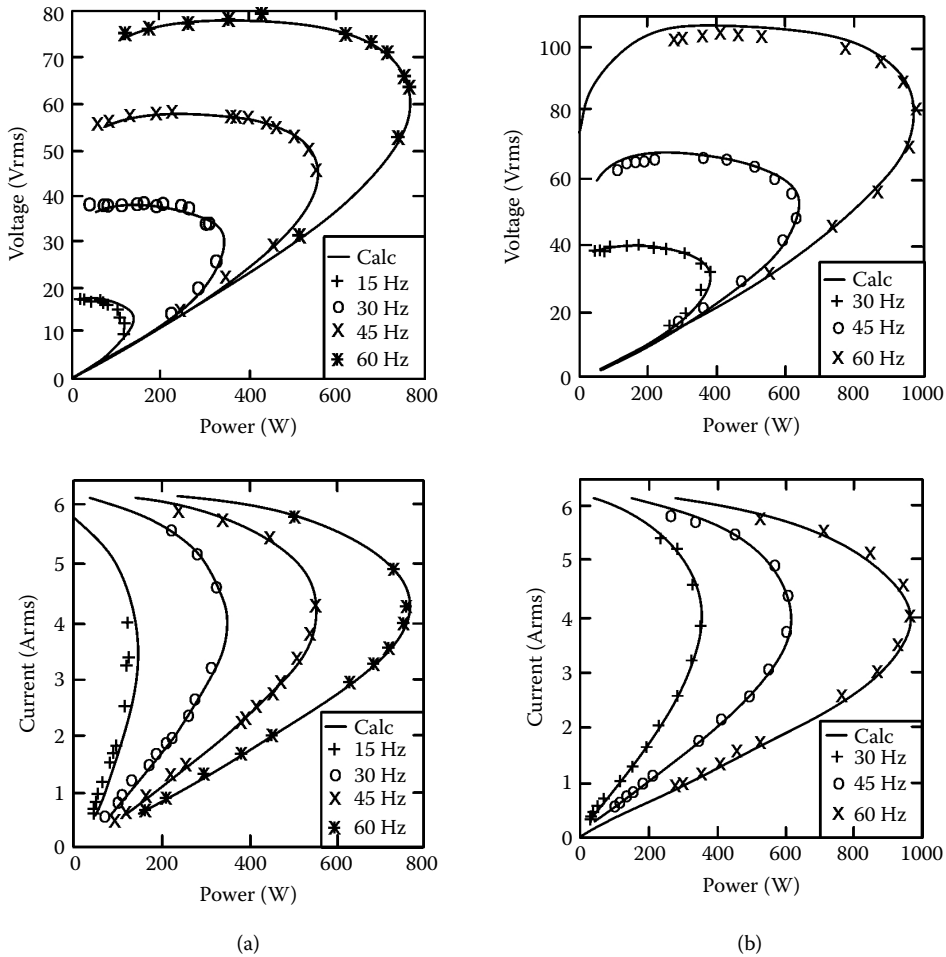


FIGURE 10.28 Interior permanent magnet synchronous generator (IPM-SG) measured and calculated voltage/power: (a) without shunt capacitors and (b) with shunt capacitors.

When magnetic saturation is considered, an iterative solution is required to account for the curve-fitted $L_d(i_s), L_q(i_s), \Psi_{PM1}(i_s)$ nonlinear functions [10].

10.7 Circuit Model of PMSG with Diode Rectifier Load

Diode rectifiers are used in many PMSG applications both for DC loads with or without battery backup or as a first stage in dual-stage AC–AC pulse-width modulated (PWM) converters with constant frequency and voltage output, for variable speed.

Let us consider here the case of diode rectifier filter and DC load (Figure 10.29).

As the d – q circuit model of the PMSG with shunt capacitors was developed in a previous section, here we will add the diode rectifier filter DC load equations.

The input and output relationships for lossless diode rectifier based on the existence function model [10] is as follows:

$$\begin{aligned} V_{dc} &= V_a \cdot S_{aV} + V_b \cdot S_{bV} + V_c \cdot S_{cV} \\ I_{aL} &= I_{dc} \cdot S_a; \quad I_{bL} = I_{dc} \cdot S_b; \quad I_{cL} = I_{dc} \cdot S_c \end{aligned} \tag{10.53}$$

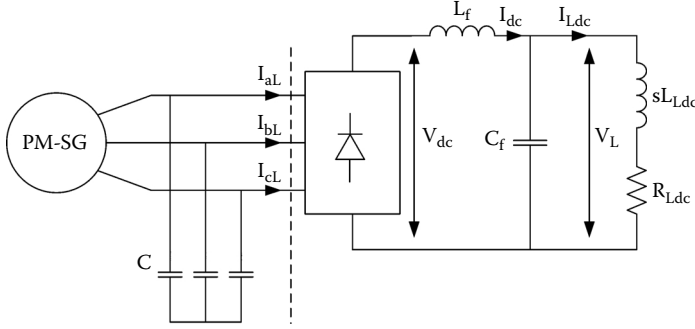


FIGURE 10.29 Permanent magnet synchronous generator (PMSG) with shunt capacitors, diode rectifier, filter, and load.

The diode rectifier voltage and current switching functions are

$$S_{a,b,c} = \frac{2\sqrt{3}}{\pi} \cos\left(\omega_r t - (i-1)\frac{2\pi}{3}\right); \quad i = 1, 2, 3$$

$$S_{a,b,c,V} = \frac{2\sqrt{3}}{\pi} \cos\left(\frac{\mu}{2}\right) \cdot \cos\left(\omega_r t - (i-1)\frac{2\pi}{3}\right); \quad i = 1, 2, 3$$
(10.54)

For steady state [20],

$$\cos\mu = 1 - \frac{2\omega_r L_c I_{dc}}{V_{LL}\sqrt{2}} \quad V_{LL} - \text{line voltage (RMS) value}$$
(10.55)

The commutation angle μ increases with the rectified current I_{dc} and with increasing machine commutation inductance. In the absence of a damper cage on the rotor, $L_c \approx (L_d + L_q)/2$. So, from the point of view of lowering the voltage drop along the diode rectifier due to machine inductances, it is beneficial to place a strong damper cage on the PM rotor. The maximum ideal value of μ should be less than $\frac{\pi}{3}$ (60°).

Approximately, the rectified DC voltage V_{dc} is related to machine line voltage [20], under steady state:

$$V_{dc} = V_{LL} \frac{3\sqrt{2}}{\pi} - \frac{3\omega_r L_c I}{\pi}$$
(10.56)

As the load was replaced by the diode rectifier, the shunt capacitor equations in d - q coordinates become as follows [19]:

$$sV_d = \frac{1}{C} I_d + \omega_r \cdot V_q$$

$$sV_q = \frac{1}{C} \left(I_q - \frac{2\sqrt{3}}{\pi} I_{dc} \right) - \omega_r \cdot V_d$$

$$V_{dc} = V_q \frac{2\sqrt{3}}{\pi} \cos\left(\frac{\mu}{2}\right)$$
(10.57)

The DC filter and load equations are added:

$$sI_{dc} = \frac{1}{L_f} (V_{dc} - V_L)$$

$$sV_L = \frac{1}{C_f} (I_{dc} - I_L)$$

$$sI_L = \frac{1}{L_{Ldc}} (V_L - R_{Ldc} I_L)$$
(10.58)

Under steady state, $s = 0$; thus, all time derivatives are zero. With the voltage drop in the diode rectifier neglected, the load resistance R_{Ldc} may be seen as a star-connected phase AC resistance $R_L = \frac{\pi^2}{18} R_{Ldc}$ in parallel with the shunt capacitors at PMSG terminals as follows:

$$3R_L \left(\frac{I_{ph}^{peak}}{\sqrt{2}} \right)^2 = R_{Ldc} \cdot I_{dc}^2 \quad (10.59)$$

$$I_{ph}^{peak} \approx \frac{2\sqrt{3}}{\pi} I_{dc}$$

Due to diode commutation of machine commutation inductances (L_c), the DC load voltage decreases notably with load [19], especially if $L_c = (L_d + L_q)/2$ (no damper cage) when the speed (frequency) is large, as in automotive applications.

With a more complete representation of a diode rectifier, the generator actual output voltage and current waveforms may be obtained [19] (Figure 10.30a and Figure 10.30b).

The presence of shunt capacitors may generate a kind of resonance phenomena with quasi-periodic oscillations with bounded dynamics in the generator line voltage and current, especially for light load.

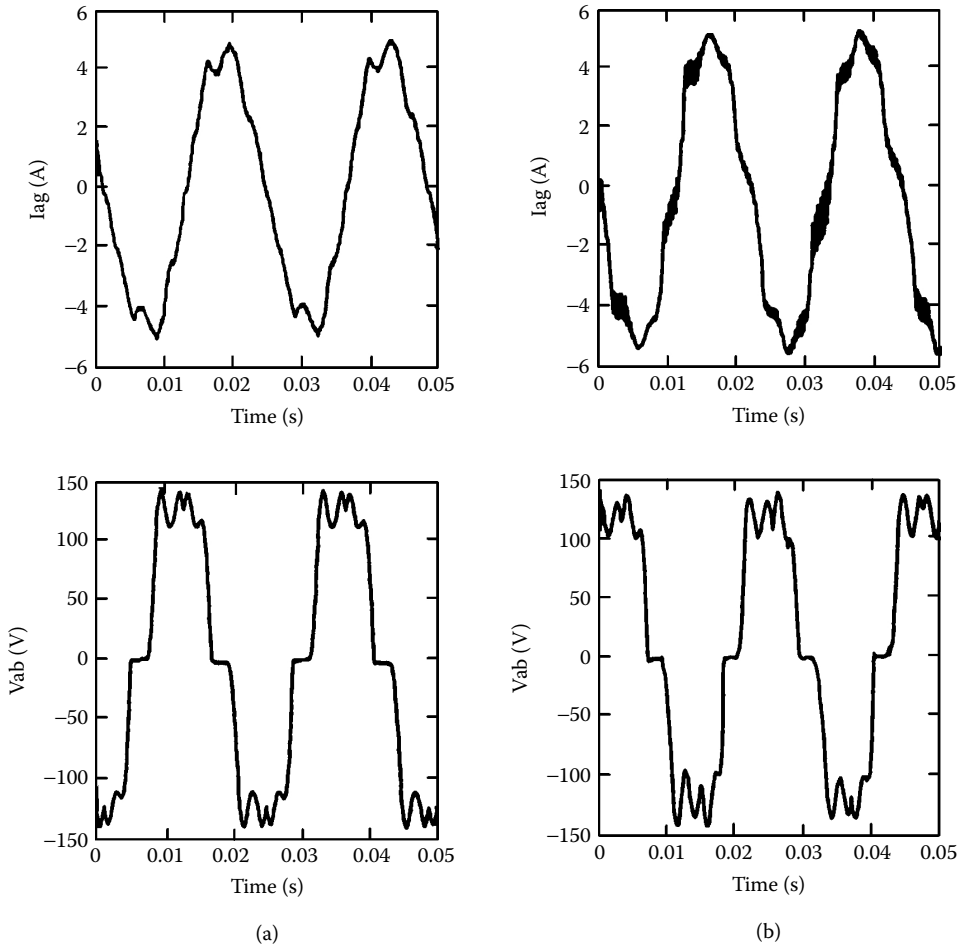


FIGURE 10.30 (a) Simulated and (b) measured generator phase current and line voltage with shunt capacitor, diode rectifier, and direct current (DC) load: $R_L = 22 \, \Omega$, $n = 1350$ rpm.

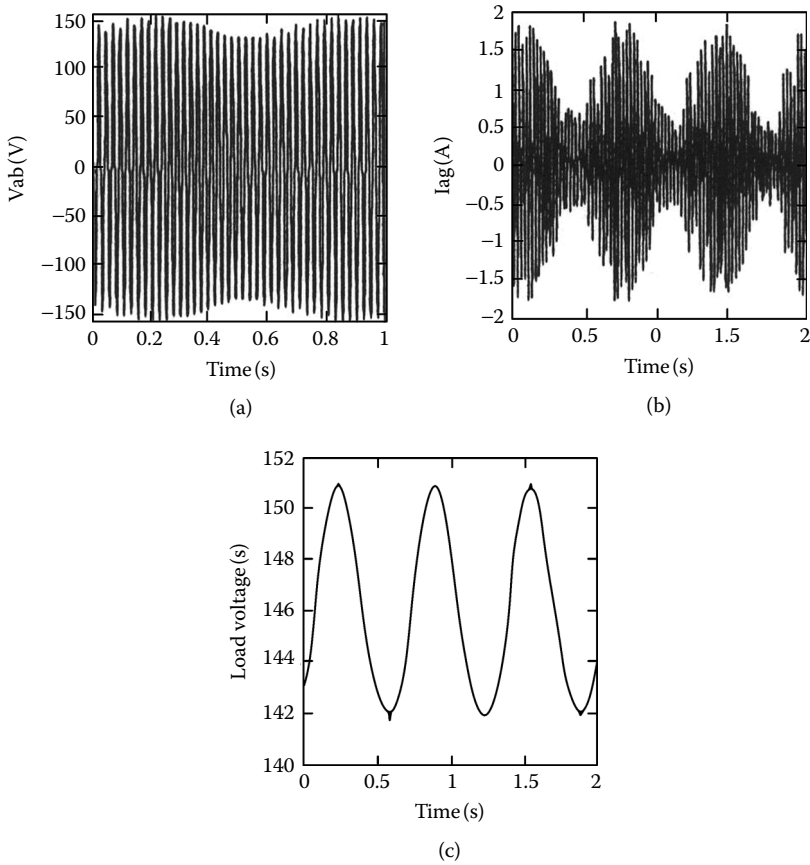


FIGURE 10.31 Oscillatory dynamics of permanent magnet synchronous generator (PMSG) with shunt capacitors and diode rectifier load: (a) generator line voltage, V_{ab} ; (b) generator phase current, i_a ; and (c) direct current (DC) load voltage.

These oscillations are visible as amplitude variations in generator line voltage and phase current and in the DC load voltage (Figure 10.31a through Figure 10.31c) [19]. Special active or passive measures have to be taken to eliminate these oscillations.

10.8 Utilization of Third Harmonic for PMSG with Diode Rectifiers

A trapezoidal AC voltage (emf) is often suggested for increased output for PMSG with rectifier DC loads. Two different rectifier topologies, as illustrated in Figure 10.32a and Figure 10.32b, are of interest.

For a purely sinusoidal system with n phases, neglecting the voltage drop in the rectifier, the ideal relationship between the AC and DC variables is as follows [16]:

$$V_{dc} = \sum_{i=1}^n V_{di} = \frac{2n}{\pi} V_1; \quad V_1 - \text{peak phase voltage} \quad (10.60)$$

$$P_{dc} = P_{ac} = V_{dc} I_{dc} = \frac{2n}{\pi} V_1 \cdot I_{ac_{rms}}; \quad n - \text{number of phases}$$

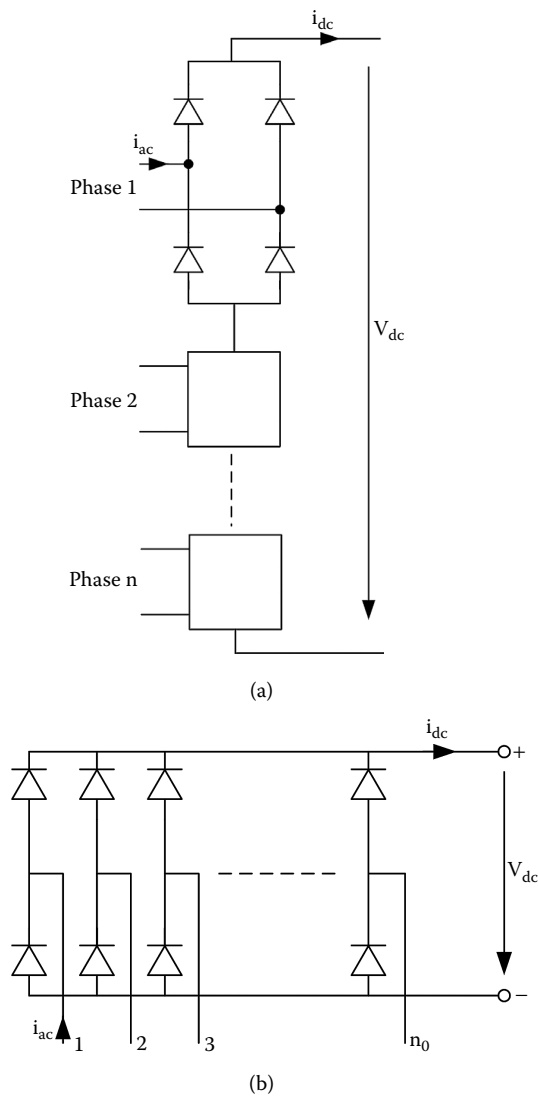


FIGURE 10.32 (a) Series (SRS) rectification and (b) polyphase (PRS) rectification.

for the series rectification, and

$$\begin{aligned} V_{dc} &= \frac{2n}{\pi} \sin \frac{\pi}{n} V_1 \\ P_{dc} &= \frac{\sqrt{2}(n)^{3/2}}{\pi} \sin \frac{\pi}{n} I_{ac_{rms}} \cdot V_1 \end{aligned} \tag{10.61}$$

for polyphase bridge rectification.
The situation changes notably when a third harmonic component $k_h V_1$ is present in the phase emfs of the PMSG.

For series rectification, simply,

$$V_{dc} = \frac{2n}{\pi} \left(1 + \frac{K_h}{3} \right) V_1 \quad (10.62)$$

$$P_{dc} = V_{dc} I_{dc} = \frac{2n}{\pi} \left(1 + \frac{K_h}{3} \right) I_{ac_{rms}} V_1$$

For the polyphase rectification, the third emf harmonic leads to multiple commutations, and thus, the DC output formulas are more involved [16] only for the voltage:

$$I_{dc} \approx \sqrt{\frac{n}{2}} I_{ac_{rms}} \quad (10.63)$$

There are basically five multiple commutation modes for polyphase rectification 0, 1, 2, 3, and 4 [16]:

$$\begin{aligned} V_{dc0} &= \frac{2n}{\pi} \left[\sin \frac{\pi}{n} - \frac{k_h}{3} \sin \left(\frac{3\pi}{n} \right) \right] \cdot V_1 \\ V_{dc1} &= \frac{2n}{\pi} \left\{ \sin \theta - \sin \left(\theta - \frac{2\pi}{n} \right) - \sin \frac{\pi}{n} - \frac{k_h}{3} \left[\sin \frac{3\pi}{n} + \sin 3\theta - \sin \left(3\theta - \frac{6\pi}{n} \right) \right] \right\} \cdot V_1 \\ V_{dc2} &= \frac{2n}{\pi} \left\{ \sin \frac{2\pi}{n} - \sin \frac{\pi}{n} + \frac{k_h}{3} \left[\sin \frac{3\pi}{n} - \sin \frac{6\pi}{n} \right] \right\} \cdot V_1 \\ V_{dc3} &= \frac{2n}{\pi} \left\{ (2 \cos \theta - 1) \sin \frac{2\pi}{n} - \sin \frac{\pi}{n} + \frac{k_h}{3} \left[\sin \frac{3\pi}{n} - (2 \cos 3\theta - 1) \sin \frac{6\pi}{n} \right] \right\} \cdot V_1 \\ V_{dc4} &= \frac{2n}{\pi} \left\{ \sin \frac{3\pi}{n} - \sin \frac{2\pi}{n} + \frac{k_h}{3} \left[\sin \frac{6\pi}{n} - \sin \frac{9\pi}{n} \right] \right\} \cdot V_1 \end{aligned} \quad (10.64)$$

with θ from

$$\sin \left(\theta - \frac{\pi}{n} \right) \cdot \sin \frac{\pi}{n} = k_h \sin \left(2\theta - \frac{3\pi}{n} \right) \cdot \sin \frac{3\pi}{n}; \quad \frac{\pi}{n} < \theta < \delta; \quad n > 5 \quad (10.65)$$

The angle δ is the offset angle of peak phase voltage value with respect to $\pi/2$. It should be noticed that a three-phase system does not show crossover (multiple commutations). A six-phase system will exhibit crossovers.

The power gain due to the presence of a third harmonic $k_h V_1$ in the PMSG phase emf is as follows [16]:

$$P_{gain}^{peak} = \frac{1 + k_h/3}{\cos \delta - k_h \cos 3\delta} \quad (10.66)$$

for constant peak voltage, and

$$P_{gain}^{rms} = \frac{1 + k_h/3}{\sqrt{1 + k_h^2}} \quad (10.67)$$

for constant RMS voltage.

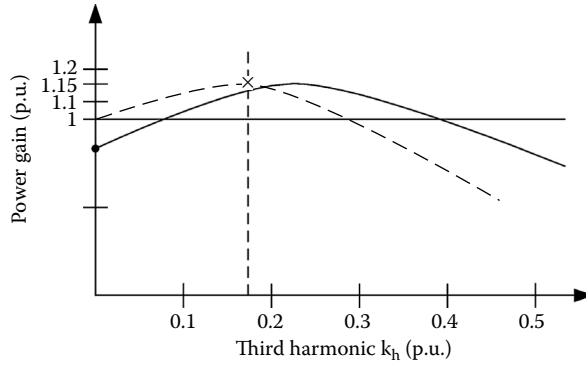


FIGURE 10.33 Power gain for three-phase rectification for constant peak voltage. The solid line represents series rectification and the broken line represents three-phase bridge rectification.

For constant peak voltage, the majority of additional power is due to increases in the fundamental, while for constant RMS voltage, it is due to increased third harmonic.

The highest power gain is obtained with a three-phase system for constant peak voltage, and typical results are shown in Figure 10.33 [16]. In the three-phase rectification, a fourth diode leg is added for the null to allow for tapping the third harmonic power.

A maximum of about 15% power gain may be obtained with both rectification methods at distinct levels of third harmonic (k_h) (Figure 10.33).

As already alluded to, the third harmonic presence in the emf (that is, trapezoidal emf) leads to additional core losses. Still, the power gain generally offsets this loss increase, and the efficiency stays constant or even increases by 1 to 2% for a 15% power gain.

Note that multiphase (nine phases, for example) PMSG windings, in groups of three, may be used to build a combination of series/polyphase rectification connections, in order to reduce DC output voltage pulsations (in amplitude and frequency) and thus allow for a smaller capacitor filter in the DC voltage link.

The presence of DC voltage pulsations and DC link capacitor filter were not considered in this section.

10.9 Autonomous PMSGs with Controlled Constant Speed and AC Load

Thermal-engine-driven autonomous PMSGs in the tens to thousands of kilowatts may be controlled for constant speed to deliver constant voltage and frequency for AC loads at moderate generator system costs. To reduce generator control system costs, fixed capacitors are connected at generator terminals. They are accompanied by thyristor-controlled three-phase inductances (Figure 10.34). For 400 Hz outputs, typical for aircraft, the size of inductors and capacitors is smaller than for 50 (60) Hz outputs.

In principle, a frequency regulator acts upon the fuel admission valve to keep the frequency (speed) constant. A voltage regulator controls the ignition angle β in the thyristor-controlled reactor and fixed capacitor voltampere reactive (VAR) compensator, to keep the voltage dynamically constant with load.

The inductor current i_{Lc} [21] is as follows:

$$i_{Lc} = \frac{V_g}{\omega L_c} \frac{(2\beta - \sin 2\beta)}{\pi} \quad (10.68)$$

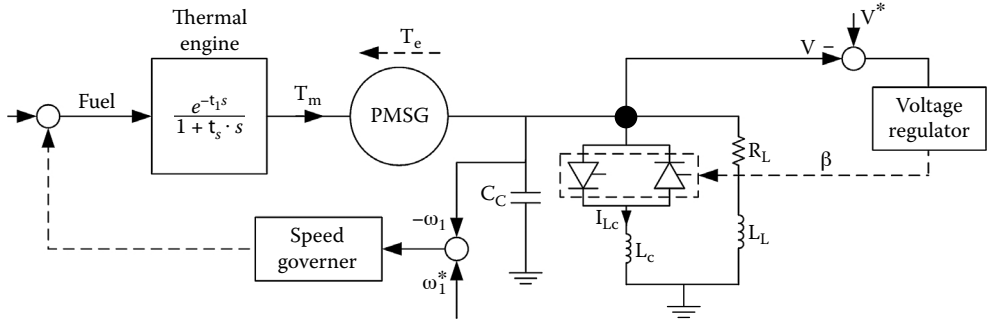


FIGURE 10.34 Permanent magnet synchronous generator (PMSG) with constant frequency (speed) and alternating current (AC) output voltage regulation.

The equivalent (variable) inductance is a function of β :

$$L_{Lc} = \frac{\pi L_c}{2\beta - \sin 2\beta} \quad (10.69)$$

To investigate the steady state and transients of the system in Figure 10.34, the d - q model is used. For steady state, the d - q model preserves the phase capacitance C , inductance L_c , and impedance Z_L along d - q axes, as the machine phase resistance and synchronous inductance are preserved.

The capacitor I_{C_c} , inductor I_{L_c} , generator I_s , and load I_L currents are visible in Figure 10.35a. For any given value of load (L_L , R_L), capacitance C_c , inductance $L_c(\beta)$, frequency, and machine parameters, the equivalent circuits or the vector diagram produce the four above-mentioned currents (Figure 10.35b). Then, the voltages V_d and V_q are obtained, and the respective active and reactive powers are calculated [21]:

$$\begin{aligned} I_d &= I_L \sin(\phi_L + \delta_V) - (I_{C_c} - I_{L_c}) \cos \delta_V \\ I_q &= I_L \cos(\phi_L + \delta_V) + (I_{C_c} - I_{L_c}) \sin \delta_V \end{aligned} \quad (10.70)$$

$$\begin{aligned} V_s &= \sqrt{V_d^2 + V_q^2} = \frac{E_0}{\cos \delta_V [1 + X_d(K_{x_L} - X_{C_c}) + R_s K_{r_L} + K_{\delta_V}]} \\ \tan \delta_V &= \frac{X_q K_{r_L} - R_s (X_{C_c} - K_{x_L})}{1 + X_q (K_{x_L} - X_{C_c}) + R_s K_{r_L}} \end{aligned} \quad (10.71)$$

with

$$\begin{aligned} K_{r_L} &= \frac{R_L}{Z_L^2}; \quad K_{x_L} = \frac{X_L}{Z_L^2}; \quad Z_L = \sqrt{R_L^2 + X_L^2} \\ X_{C_c} &= \frac{1}{X_c} - \frac{1}{X_{Lc}}; \quad K_{\delta_V} = [X_d K_{r_L} - R_s (K_{x_L} - X_{C_c})] \tan \delta_V \end{aligned}$$

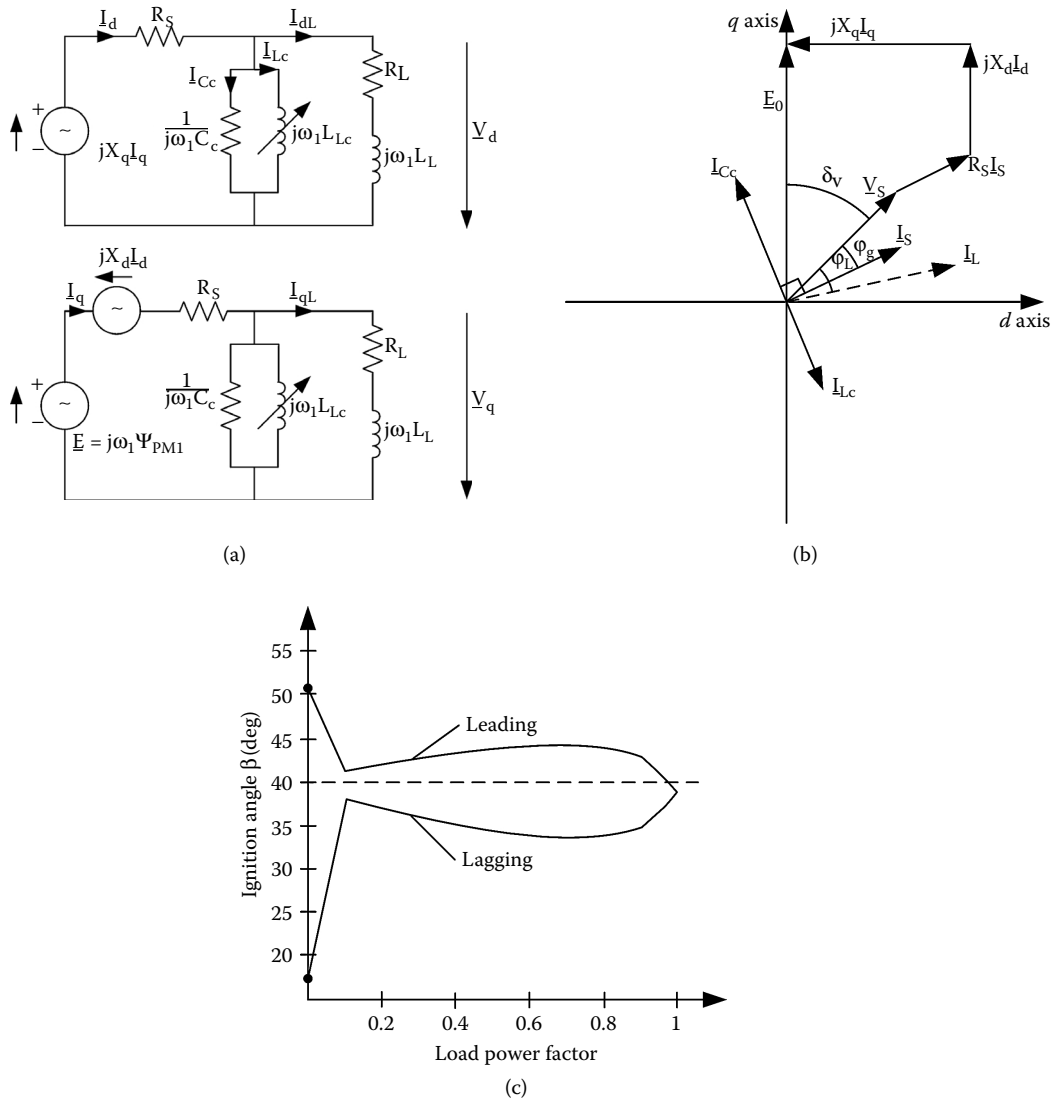


FIGURE 10.35 Steady-state equivalent circuits in phasor variables decomposed in d - q components: (a) the circuits, (b) the vector diagram, and (c) typical β vs. load power factor at constant load current.

Once the terminal V_s and the voltage power angle δ_v are calculated, all the other variables (currents, powers) are straightforwardly determined as follows:

$$I_L = \sqrt{I_{dL}^2 + I_{qL}^2}; \quad I_{Lc} = \frac{V_s}{\omega_1 L_{Lc}}; \quad I_{Cc} = V_s \cdot \omega \cdot C_c$$
$$I_s = \sqrt{I_d^2 + I_q^2} = \frac{V_s}{Z_L}$$

(10.72)

Given the range of load impedance Z_L and its power factor angle ϕ_L variation range, the required capacitor and inductance that provide for constant terminal voltage for constant speed may be

determined. When defining the variable inductance value L_C , the angle β is from Equation 10.69, and thus, the voltage drop compensation conditions may be investigated as a function of the ignition angle β .

Figure 10.35c shows a typical angle β dependence on load power factor at rated load current and voltage [21].

For all load power factor angles (except for $\cos\phi_L = 1$), there are two values of ignition angle — one for leading power factor and one for lagging power factor.

While the design of C_C , L_C for steady state is the first step, the investigation of transients and control aspects is crucial.

The d - q model, in the rotor coordinates, is used for the scope. For the sake of completeness, the system equations are given below — first for the load and the VAR compensator:

$$\begin{aligned}
 sL_d &= \left(\frac{V_d - R_L i_{dL}}{L_L} \right) + \omega_r i_{Lq} \\
 sL_q &= \left(\frac{V_q - R_L i_{qL}}{L_L} \right) - \omega_r i_{Ld} \\
 sV_d &= \frac{i_{Cd}}{C} + \omega_r V_q \\
 sV_q &= \frac{i_{Cq}}{C} - \omega_r V_d \\
 si_{d_{Lc}} &= \frac{V_d}{L_{Lc}} + \omega_r i_{q_{Lc}} \\
 si_{q_{Lc}} &= \frac{V_q}{L_{Lc}} - \omega_r i_{d_{Lc}}
 \end{aligned} \tag{10.73}$$

The d - q model of PMSG is described by Equation 10.52, with i_d , i_q as dummy variables. The motion equation has to be added:

$$\frac{J}{p_1} \frac{d\omega_r}{dt} = T_m - T_e; \quad T_e = \frac{3}{2} p_1 (\Psi_d i_q - \Psi_q i_d) \tag{10.74}$$

The above equations are first linearized and so is the thermal engine (say diesel engine) equation:

$$\tau_s \cdot s \cdot \Delta T_m + \Delta T_m = \Delta F(t - \tau_1) \tag{10.75}$$

τ_1 is the engine dead time (sampling time, approximately), and τ_s is the electrohydraulic actuator time constant.

While independent speed (frequency) and voltage regulators may be adopted, more advanced state feedback, or output feedback, controls were introduced for the scope, with good results [21]. Typical transients with optimal state feedback control are shown in Figure 10.36, based on the results from Reference 21.

A notable short-lived reduction in the terminal voltage is noticed. However, this event is followed by a fast recovery. The fuel rate increases steadily as the load increases. The speed deviation transients are small, while the thyristor ignition angle varies markedly to accommodate the new, larger, load current at about the same load power factor.

Given the numerous nonlinearities in the system, nonlinear (such as variable structure) control may be applied to further reduce the voltage surges during transients.

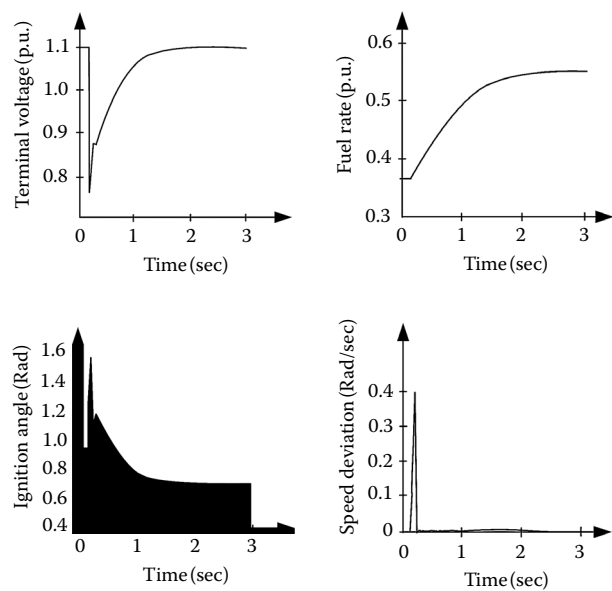


FIGURE 10.36 State feedback controller response when the load impedance varies suddenly from $(1.8 + j1.35)$ P.U. to $(1.2 + j0.9)$ P.U.

10.10 Grid-Connected Variable-Speed PMSG System

Variable-speed use is good for extracting more prime-mover power (as in wind turbines) or for providing optimum efficiency for the prime mover by increasing its speed with power. Variable speed also allows for a more flexible generator system. For wind turbines, a battery may be added to store the extra wind energy that is not momentarily needed for the existing loads (or local power grid).

There are two main ways to handle the necessity of stable (constant) DC link voltage at variable speed as illustrated in Figure 10.37a and Figure 10.37b.

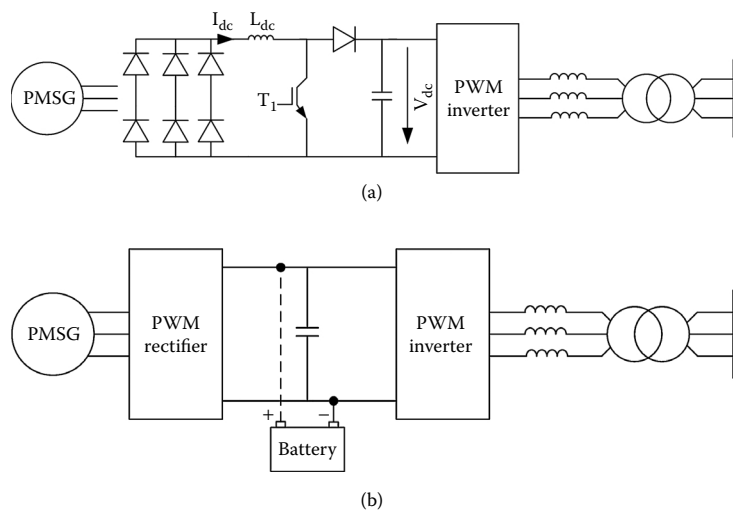


FIGURE 10.37 Variable-speed permanent magnet synchronous generator (PMSG) system with constant output voltage and frequency: (a) with diode rectifier and boost direct current (DC)–DC converter and (b) with pulse-width modulator (PWM) rectifier.

For the case of the diode rectifier, the generator is designed with terminal voltage at maximum speed to allow for a small voltage boost in the DC–DC converter for all situations.

Full swing boost is required at minimum speed.

10.10.1 The Diode Rectifier and Boost DC–DC Converter Case

In the absence of a rotor cage, the d – q (space-vector) model of PMSG (after Equation 10.52), for the generator mode association of signs, is as follows:

$$\begin{aligned}\bar{V}_s &= -\bar{i}_s R_s - j\omega_r \bar{\Psi}_s \\ \bar{\Psi}_s &= \Psi_d + j\Psi_q; \quad \bar{i}_s = i_d + j i_q; \quad \bar{V}_s = V_d + j V_q\end{aligned}\quad (10.76)$$

$$\bar{\Psi}_s = \Psi_{PM} + L_d i_d + j L_q i_q \quad (10.77)$$

As for steady state ($s = 0$), the diode rectifier imposes almost unity power factor at the fundamental frequency:

$$Q_1 = 0 = \frac{3}{2} (V_d i_q - V_q i_d) \quad (10.78)$$

Consequently, from Equation 10.76 through Equation 10.78, the following condition is obtained:

$$L_q i_q^2 + L_d i_d^2 - \Psi_{PM} i_d = 0 \quad (10.79)$$

In this writing, $i_d > 0$.

The electromagnetic torque T_e is as follows:

$$T_e = \frac{3}{2} p_1 (\Psi_{PM} - (L_d - L_q) i_d) i_q; \quad L_d < L_q; \quad i_d > 0 \quad (10.80)$$

with $L_d = L_q = L_s$, Equation 10.79, when used in Equation 10.80, yields the following:

$$\frac{\partial T_e}{\partial i_d} = \frac{\partial}{\partial i_d} \left(\frac{3}{2} p_1 \Psi_{PM} \sqrt{\frac{i_d \Psi_{PM} - L_s i_d^2}{L_s}} \right) = 0 \quad (10.81)$$

with the following solution:

$$i_{dk} = \frac{\Psi_{PM}}{2L_s} \quad (10.82)$$

So, the maximum torque T_{ek} is

$$(T_{ek})_{\cos \varphi_1=1} = \frac{3}{2} p_1 \frac{\Psi_{PM}^2}{2L_s}; \quad L_s = L_m + L_{sl} \quad (10.83)$$

The larger the stator leakage inductance L_{sl} , the lower the maximum torque at unity power factor. Also, for maximum torque, from Equation 10.79 and Equation 10.82,

$$i_{qk} = i_{dk} = \frac{\Psi_{PM}}{2L_s} \quad (10.84)$$

But, the condition in Equation 10.84 corresponds, in general, to maximum torque/current, which is to be expected, as the power factor was taken as unity, for the diode rectifier case. With base torque T_b defined as follows,

$$T_b = \frac{P_b}{\omega_{rb}} \cdot p_1 \quad (10.85)$$

the maximum torque is

$$\frac{(T_{ek})_{\cos \phi_1=1}}{T_b} = \frac{1}{2X_s(p.u.)} \quad (10.86)$$

As expected, the lower the X_s (in P.U.), the larger the peak torque. Reducing the stator leakage inductance becomes a very important design task.

The control of power flow to the power grid depends on both the DC link voltage level (controlled by the PWM inverter on the power grid side) or on DC–DC boost converter output DC voltage.

For a wind turbine, the maximum wind power extraction occurs for a given $\omega_r(T_m^*)$ reference curve up to maximum power value, when stall regulation occurs. The DC current may then be regulated to prescribe the modulation index of the IGBT in the DC–DC boost converter (Figure 10.38).

To limit the DC–DC converter chopping frequency, no external means are required, as the large commutation inductance (L_s) of the generator is connected in series with L_{dc} during on states of T_1 [22,23].

As already mentioned in this section, an active machine end (PWM) rectifier may replace the diode rectifier plus the boost DC–DC converter. In this case, the voltage at the machine terminals, at lower speeds, may be lower than the constant-kept capacitor voltage in the DC link by the action of the grid-side PWM converter. In essence, the capacitor in the DC link provides, in a controlled manner, the required reactive power injection into the machine in addition to the commutation reactive power in the rectifier.

The PMSG may produce more torque than for the case of the diode rectifier but does this also for higher stator currents, as required by the apparent power of the generator.

The back-to-back PWM converter system with DC capacitor link may provide both variable speed operation and reactive power controlled delivery at constant AC output voltage and frequency.

The machine-side converter is torque (speed) controlled along the q axis and is constant flux controlled along the d axis field-oriented control (FOC) or in direct torque and flux control (DTFC) schemes.

On the other hand, the load-side converter is controlled differently for stand-alone operation.

In power grid operation, the grid-side PWM converter keeps the DC link voltage constant and delivers a certain amount of reactive power, as desired. With an adequate design for the DC link capacitor, up to ± 100 reactive and active power control is feasible with PMSGs at a lower cost of the capacitor in comparison with the induction generator with cage rotor.

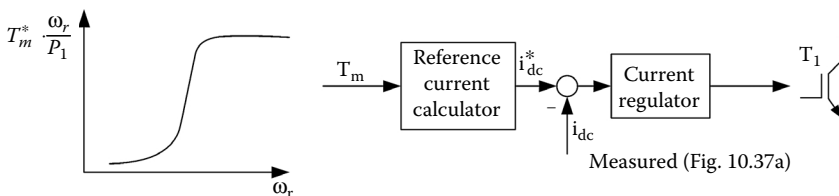


FIGURE 10.38 Boost direct current (DC)–DC converter control for a wind-turbine permanent magnet synchronous generator (PMSG).

For stand-alone loads, an L-C power filter is mandatory between the load-side PWM inverter and the loads. Now the output voltage frequency is set at a constant value, while the AC output voltage amplitude, after the output filter, is controlled through the DC link capacitor voltage close-loop control via, again, the load-side converter.

The PMSG, once in rotation, with the machine-side converter idle, charges the DC link capacitor through its diodes to offer conditions for self-excitation of the system on no load.

A backup battery and a solar panel may work together with the PMSG systems with wind or hydro-turbines to compensate for the limited availability, dependent on time of day, of wind (hydro) energy.

10.11 The PM Genset with Multiple Outputs

Medium-power (tens to hundreds of kilowatts) gensets were traditionally designed to be rugged and cost effective, though heavier, for special applications. In essence, their speed, when driven by a diesel engine, is in the 1800 to 3600 rpm range; while for gas turbines, it may be in the tens of thousands-revolutions per minute.

Gensets may be used as auxiliary power sources on ground (for aircraft land service), as standby power supply units, or on trucks, vessel, or aircraft as medium-power auxiliary power sources. The multiple output possibility stems from the capability of the PWM inverter on the load side to be controlled for various voltages, frequencies, and numbers of phase outputs (Figure 10.39) [24].

The diode rectifier, plus the boost DC-DC converter, is used to produce constant DC link voltage over a wide speed range (up to about 2/1 range).

The bidirectional buck-boost DC-DC converter is used to supplement the $28 V_{dc}$ loads and recharge the $28 V_{dc}$ backup battery, and, in addition, to provide a DC voltage boost whenever necessary during load or speed transients.

The standard PWM 6 IGBT voltage source inverter is controlled, on a menu basis, to produce AC output at 50 or 60 Hz, or 400 Hz (for example), and with single-phase, dual-phase, or three-phase attributes, according to needs.

Variable speed is used to reduce fuel consumption (Figure 10.40) [24] or to reduce thermal engine pollution, by reducing speed when required power is reduced.

As expected, there is minimum fuel consumption for almost constant power at a certain speed (2900 rpm in Figure 10.40).

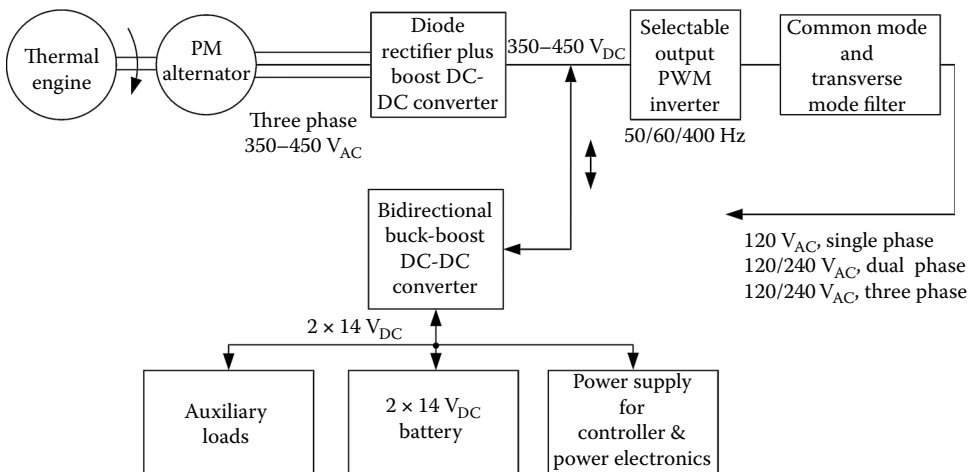


FIGURE 10.39 A permanent magnet (PM) Genset with multiple outputs.

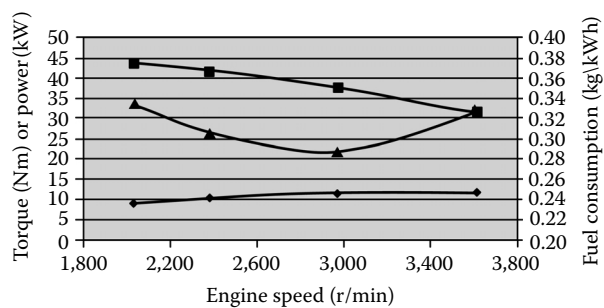


FIGURE 10.40 Peak torque (▲), power (■), and fuel consumption (◆) of a diesel engine permanent magnet (PM) Genset.

The machine-side diode rectifier is standard for a three-phase PMSG, but the boost DC–DC converter may take various configurations. A representative one is shown in Figure 10.41 [24].

Given the rather large value of generator commutation (subtransient, almost synchronous) inductance, the latter is used for voltage boost. The accumulation of energy in the machine inductances is performed through the additional diode rectifier D_2 , which is short-circuited by the IGBT, Q_1 .

Subsequently, Q_1 is disengaged, and the higher voltage obtained at the output of the main diode rectifier D_1 charges the capacitor C_1 in the DC link to boost the DC link voltage. The snubber C_2 – R_1 limits the di/dt spikes at Q_1 when the latter is disengaged. The alternator and load neutrals are connected to the center of two capacitors C_3 in the DC link (Figure 10.42). The neutral connection allows for unbalanced load currents to flow through the PMSG in parallel with the capacitors C_3 and, thus, eliminates the necessity that the latter have low fundamental frequency impedance.

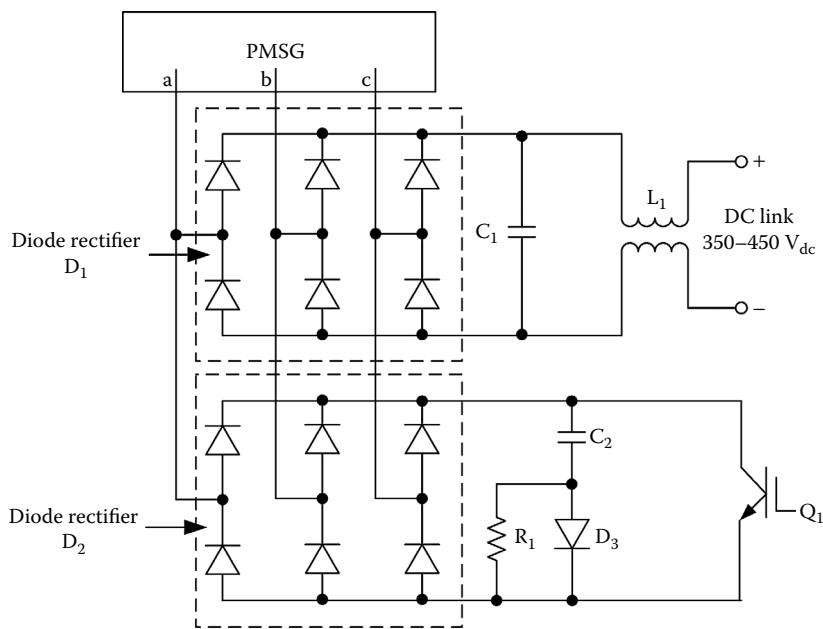


FIGURE 10.41 Diode rectifier plus boost direct current (DC)–DC converter.

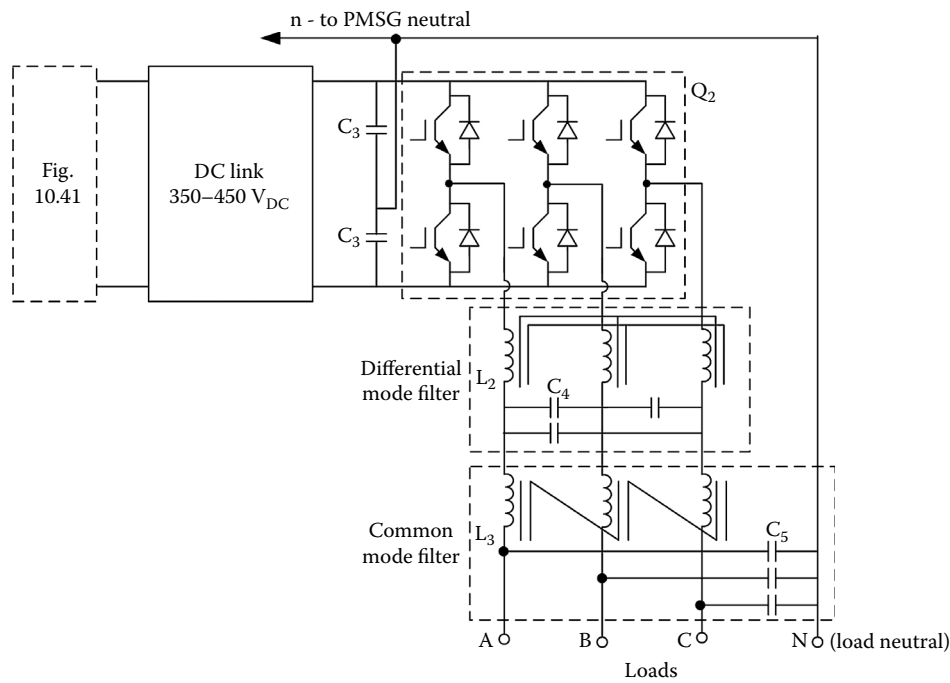


FIGURE 10.42 The permanent magnet synchronous generator (PMSG) inverter and filters.

The inductance L_1 (Figure 10.41) avoids the charging of capacitors C_3 to the half-wave rectified line to neutral voltage. Poor power factor and high crest currents in the generator are thus avoided.

The three-phase differential (L_2, C_4) and common-mode (L_3, C_5) filters are also shown in Figure 10.40 [24]. They are meant to achieve near sine wave output (load) voltage ($< 3\%$ total harmonic distortion [THD]). Line to neutral inverter output voltage pre- and post-filter output voltages shown in Figure 10.43 are fully indicative of a filter’s beneficial effect [24].

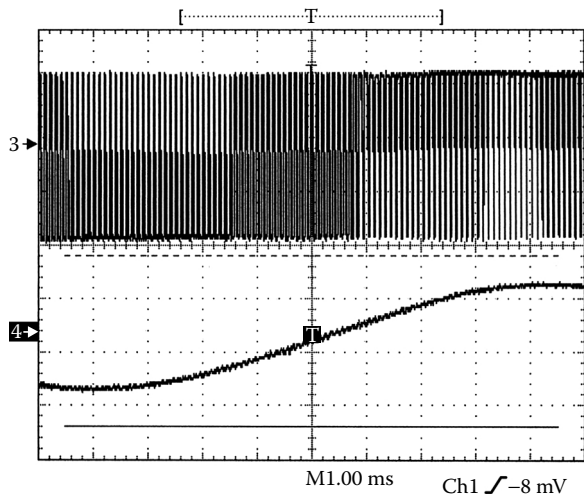


FIGURE 10.43 Pre- and post-filter line to neutral load voltage (V_{AN}).

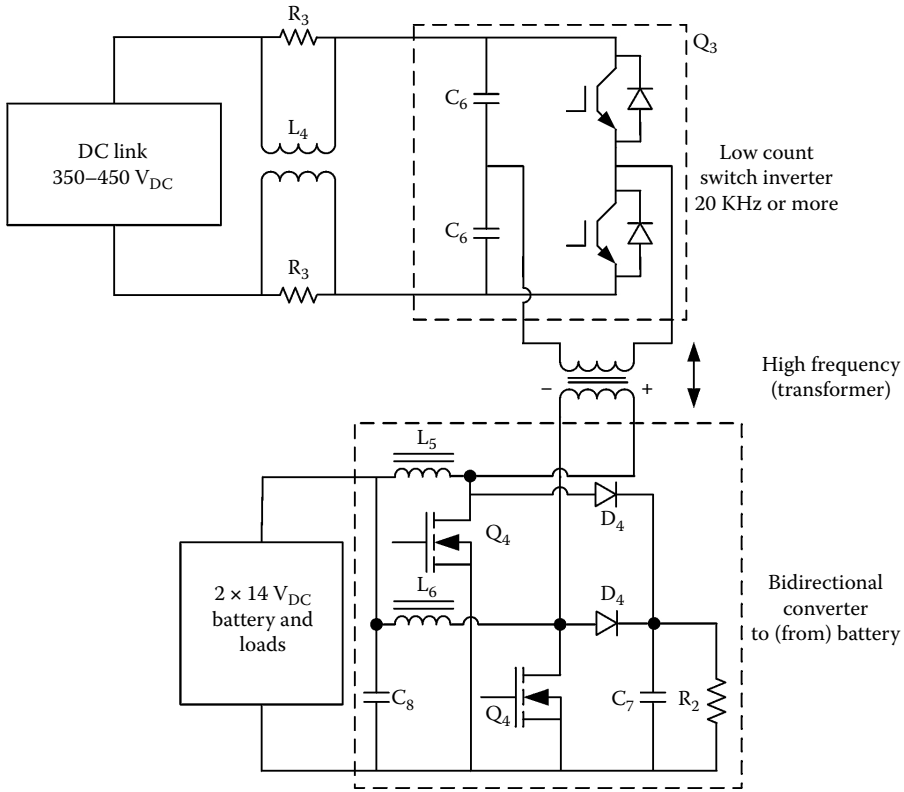


FIGURE 10.44 Bidirectional direct current (DC)–DC power converter.

The bidirectional DC–DC converter converts power from (to) 350 V_{dc} to (from) 28 V_{dc} to supply auxiliary loads and to recharge the engine-starting battery that supplies, in turn, the engine starter (a DC PM motor).

There are many competitive bidirectional buck-boost DC–DC converter configurations for such a high voltage ratio, but they all make use of a high-voltage-frequency transformer.

Such a typical configuration making use of low power switch count is shown in Figure 10.44 [24].

For high to low voltage operation modes, the DC link high voltage is first passed through the damper R_3 – L_4 to prevent oscillations between capacitors C_6 and C_3 (of the PWM three-phase converter). The capacitors C_6 make up for the missing half of the low count switch inverter/rectifier and provide a one half reduction in voltage, up front. The transformer T_1 may provide a one half to one third voltage reduction ratio. A second reduction of one half of the voltage is performed by a current doubler through the diodes D_4 and filter L_5 , L_6 . Thus, the low count switch converter Q_3 is operated at a maximum duty ratio of less than 0.85 [24].

When the DC link voltage is below 350 V_{dc} (340 V_{dc} and less), the low-to-high-voltage conversion is engaged.

When both Q_4 metal-oxide semiconductor field-effect transistors (MOSFETs) are turned on, the secondary of the transformer T_1 is short-circuited, and the inductances L_5 , L_6 are energized. When only one of them is on, the input voltage plus the energy stored in one inductor is applied to the secondary of the transformer. This, in turn, steps up the voltage that is then rectified through the diodes in the low switch count converter Q_3 . When both Q_4 are off, diodes D_4 discharge the energy stored in L_5 and L_6 to capacitor C_7 , which is discharged by the resistance R_2 .

Even under complete loss of power from the generator, the bidirectional boost DC–DC converter should be capable of holding the DC link high voltage for a few seconds. In general, the voltage boost operation mode is needed to complement the generator power short-lived current peaks and thus maintain the load voltage.

A fast load power increase is, for example, quickly accommodated before the thermal engine speed is increased to sustain it by additional, controlled fuel injection.

10.12 Super-High-Speed PM Generators: Design Issues

By super-high speeds, we mean here speeds above 15,000 rpm. Power decreases with speed to limit the rotor peripheral speed. It is up to 100 to 170 kW at 70 to 80,000 rpm and goes up to 1 to 5 megawatt (MW) at 12,000 to 18,000 rpm. Small- to medium-power super-high-speed micro-turbine-driven PM generators are envisaged for distributed power generation or on board series hybrid vehicles, and for other usage. Drastic reductions in size and weight make such generator systems very attractive. The main design features are as follows:

- Electromagnetic design (sizing)
- Mechanical design (verification)
- Thermal design

The starting point is the specifications. They depend on application, and here is an example:

- Rated electric power: $P_{en} = 80$ kW
- Rated speed: $n_n = 80,000$ rpm
- Minimum speed at rated power: $n_{min} = 60,000$ rpm
- Rated power factor: $\cos \varphi_n = 0.8$
- Number of poles: $2p_1 = 4$
- Rated phase voltage (RMS): 220 V

In the electromagnetic design (sizing), due to the severe rotor mechanical constraints, the latter should be considered first.

10.12.1 Rotor Sizing

Two main rotor configurations are to be considered — the cylindrical rotor and disk-shaped rotor (Figure 10.45a and Figure 10.45b). In general, the disk-shaped PM rotors are performing better with a larger number of poles $2p_1 > 6, 8$. Multiple disk-shaped rotors are used to increase power for a given speed. As $2p_1 = 6, 8, 12, 14, 16$, the stator fundamental frequency tends to be larger, so the maximum speed with $2p_1 = 16$ would be 15,000 rpm for $f_1 = 2.0$ kHz, and 30,000 rpm for $2p_1 = 8$. The maximum peripheral speed of the rotor, say 300 m/sec, is another limiting factor for the maximum allowable power and speed. The cylindrical rotors are used, in general, with $2p_1 = 2, 4$. Better power/volume performance is claimed with multiple disk rotor PMSGs.

As mentioned at the beginning of this chapter, the copper thin shields are crucial in reducing the rotor eddy currents produced by the space and time stator field harmonics in the PMs and in the rotor back iron. Solid back iron is acceptable with copper shields, which leads to more rugged rotors. Typical material properties for the rotor are as follows:

- Retaining ring: carbon fiber + epoxy resin: 2100 MPA
- Shaft: nonmagnetic iron: 1300 MPA
- Rotor mass: magnetic iron: 1800 MPA
- Interpolar spacer: nonmagnetic iron: 1300 MPA

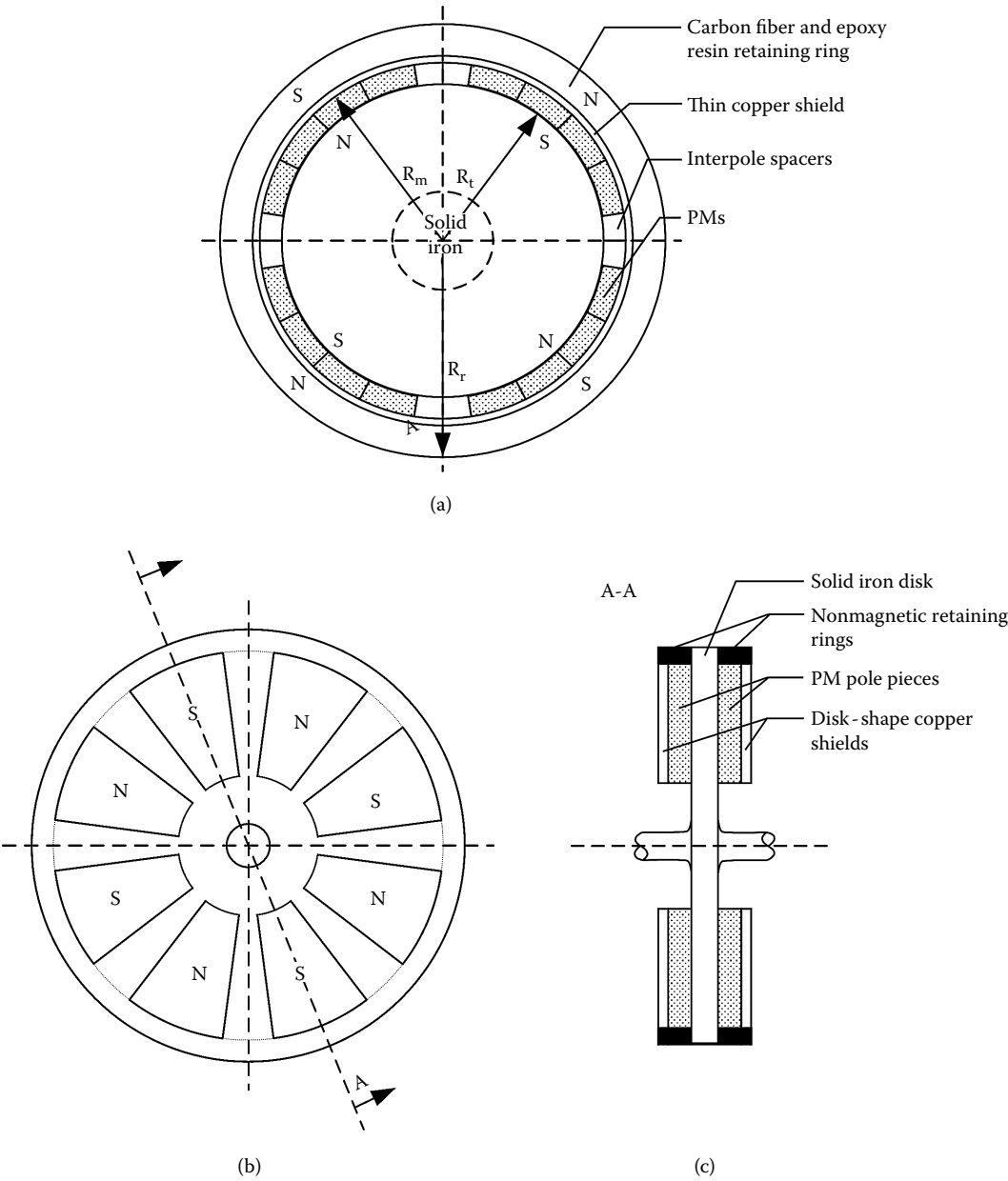


FIGURE 10.45 Super-high-speed permanent magnet (PM) rotors: (a) and (b) cylindrical and (c) disk-shaped.

To size the cylindrical rotor PMSG, the following main variables have to be determined:

- Inner stator diameter D_{is} and stack length
- Rotor geometry details and mechanical validations
- Stator sizing and thermal verification

A similar path is to be traveled for the disk rotor (axial airgap) PMSG, but with adapted (new) formulas. The analytical preliminary design has to be followed by FEM comprehensive verification and design optimization.

The specific tangential force f_{tk} in the range of 0.5 to 3 N/cm² is used here to obtain the preliminary value for stator interior diameter D_{is} , for given λ :

$$f_{tn} = T_{en} \cdot \frac{2}{D_{is}} \cdot \frac{1}{\pi \cdot D_{is} \cdot \lambda \cdot D_{is}}; \quad \lambda = l_{stack} / D_{is} \quad (10.87)$$

$$T_{en} \approx \frac{P_{en}}{\eta_n \cdot 2\pi n_n} \quad (10.88)$$

where η_n is the rated (forecasted) efficiency. For super-high-speed PMSGs, $\lambda \approx 1$ to 3.0. The rated efficiency is assigned a value of 0.9 to 0.95 at this early stage of design. Also, f_{tn} is higher when the rated torque increases.

The mechanical airgap g should be larger than usual to reduce the space harmonics of the stator current magnetic flux at the rotor copper shield outer surface caused by the stator mmf and slot-opening. This way, the losses in the copper shield will be reduced.

In general, the stator slot opening W_{es} should be the same as the radial distance h_{tg} between the stator and the PMs:

$$h_{tg} = g + b_{rr} + b_{copper} > W_{es} \quad (10.89)$$

where

$b_{rr} = 1.0$ to 3 mm retaining ring thickness

$b_{copper} = 0.5$ to 1.5 mm

$g = 1$ to 2 mm

For the same reason, whenever possible, the stator slot pitch τ_s should be of the size of the total (magnetic airgap) h_{Mg} :

$$h_{Mg} = h_{tg} + h_{PM} \approx \tau_s \quad (10.90)$$

with h_{PM} equal to the PM thickness.

In general, the PM airgap flux density B_{gPM} in the airgap is quasi-rectangular:

$$B_{gPM} \cong \frac{B_r}{(1+k_f)} \cdot \frac{h_{PM}}{h_{Mg}} \quad (10.91)$$

The fringing factor k_f accounts for the PM flux lines that close into the airgap or through stator slot necks, instead of embracing the stator slots. As h_{tg} gets larger, k_f increases. The value of $k_f = 0.3$ to 0.7 are not uncommon, but they have to be checked through FEM calculations.

In general, $B_{gPM} = 0.3$ to 0.45 T in super-high-speed PMSGs, as the fundamental frequency is in the 600 Hz to 2.5 kHz range, and thus, core losses tend to be large.

The PM flux per half a pole pitch $\Phi_p/2$ is as follows:

$$\Phi_p/2 = \Phi_{yr} = B_{gPM} \cdot \frac{\tau_{PM}}{2} \cdot l_{stack} \cdot k_{fill} \quad (10.92)$$

k_{fill} is the lamination filling factor. Even with thin (0.1 mm or so) laminations, the insulation coating should keep $k_{fill} > 0.7$ to 0.8. τ_{PM} is the PM span length on one rotor pole.

Half the PM flux/pole passes through the rotor back iron, with radial dimension h_{yr} that needs to be as follows:

$$h_{yr} \geq \frac{\Phi_{yr}}{l_{stack} \cdot B_{yc0}} \quad (10.93)$$

B_{yc0} is the rotor yoke design flux density. Generally, $B_{yc0} \cong 0.9$ to 1.2 T (here at no load) to leave room for armature reaction field on load but still maintain B_{yc} on load below 1.3 to 1.5 T.

The shaft diameter D_{shaft} is, basically,

$$D_{shaft} = D_{is} - 2h_{PMg} - 2h_{yr} \quad (10.94)$$

This value has to be checked for mechanical strength.

The rotor mechanical stress verification starts with the computation of spin-off stress σ_{max} , which contains the PM plus copper shield and its own weight contributions:

$$\sigma_{max} = \frac{F_E}{A} = \frac{\omega_r^2 (\gamma_{PM} V_M R_m + \gamma_{copper} V_{copper} (R_m + b_{copper}))}{b_{rr} l_{stack}} + \gamma_{ring} \cdot R_t^2 \cdot \omega_r^2 < \frac{\sigma_{max}}{k_{safe}} \quad (10.95)$$

The safety coefficient $k_{safe} = 2 \div 3$; R_m and R_t are visible in [Figure 10.45a](#).

The retaining ring radial elongation due to the applied stress should also be limited:

$$\Lambda = \frac{\sigma_{max}}{E} \quad (10.96)$$

E equals the modulus of elasticity of the ring.

Critical flexural speeds of the rotor also have to be checked, as gravity bends the shaft, and centrifugal and magnetic uncompensated radial forces produce mechanical radial oscillations. There may be more than one critical rotor speed below the rated (superhigh) speed. With one critical speed ω_k , the latter should be much less than the rated speed ω_n ; $\omega_k < \omega_n/4$, in general. Driving through critical speeds should be quick to avoid large vibration and noise.

10.12.2 Stator Sizing

The stator sizing means, essentially, the slotting and winding design once the stator bore diameter and stack length are already calculated.

Typically, for super-high-speed PMSGs, distributed double-layer windings with chorded coils are used, as the number of poles is limited.

Also, when the fundamental frequency goes up to 2.2 to 2.5 kHz, the skin effect winding losses require special treatment, through design, in order to maintain the skin effect losses within 30 to 35% of fundamental copper losses. Sinusoidal current control seems to be the preferred choice, and when a diode rectifier is connected at machine terminals, above 500 Hz fundamental frequency, this is the actual situation anyway, due to commutation overlapping.

Rectangular current control with trapezoidal emf may also be practical, as fundamental frequency and power go up, and thus, the limited switching frequency in the PWM converter is better suited for such six-pulse/cycle operation.

Let us suppose here sinusoidal current control with sinusoidal emf. The phase emf fundamental E_1 per current path is as follows:

$$E_1 = \pi \sqrt{2} \Phi_p W_a K_{w_1} f_1 \quad (10.97)$$

where

W_a is the number of turns per current path (in series)

K_{w_1} is the winding factor for the fundamental

f_1 is the stator frequency

With q_1 slotspole/phase, W_a (for two coils per slot) is as follows:

$$W_a = \frac{2n_c p_1 q_1}{a} \quad (10.98)$$

where

n_c is the number of turns per coil

a is the current path count

Denoting by V_1 the rated phase voltage (RMS value), the design ratio (E_1/V_1) depends on the type of machine end converter, overload, type of load, and speed range. Generally, at rated speed, $E_1/V_1 \approx 1.1$ to 1.3.

Once the winding type is chosen (two layers, chorded coils, and q_1 as given), the winding factor for the fundamental, K_{w_1} , is as follows:

$$K_{w_1} = \frac{\sin \pi/6}{q_1 \sin \pi/6q_1} \cdot \sin \left(\frac{Y}{\tau} \cdot \frac{\pi}{2} \right) \quad (10.99)$$

Consequently, from Equation 10.97, the number of turns per path W_a (with a given) can be calculated.

The rated current I_{1n} comes directly from the specifications:

$$I_{1n} = \frac{P_{en}}{(\cos \varphi_n) 3V_{1phn}} \quad (10.100)$$

The number of Ampere-turns per slot F_{1slot} is

$$F_{1slot} = 2n_c \frac{I_{1n}}{a} \quad (10.101)$$

Adopting a design current density $j_{con} = 3$ to 8 A/mm² for air cooling and higher for liquid cooling, the conductor-filled area of the slot A_{su} is as follows:

$$A_{su} = \frac{F_{1slot}}{k_{fills} \cdot j_{con}} \quad (10.102)$$

$k_{fills} \approx 0.3$ to 0.45 is the slot filling factor. Smaller values of k_{fills} have to be considered for direct cooling of conductors in slot by pipes (or hollow conductors).

The tooth width W_t is imposed by the tooth flux density B_t :

$$B_t \approx \frac{B_{gPM} \cdot \tau_s}{W_t} \leq 1.0 - 1.2 \text{ T}; \quad \tau_s = \frac{\pi D_{is}}{6 p_1 q} \quad (10.103)$$

The stator yoke height h_{ys} is about equal to the rotor yoke height h_{yr} :

$$h_{ys} \approx \frac{\Phi_p}{2 l_{stack} B_{ys}} \quad B_{ys} < (0.9 - 1.2) \text{ T} \quad (10.104)$$

At $\tau_s / W_t \approx 1.8 - 2.2$ and knowing A_{su} leads to complete sizing of the slot. Approximately, the slot useful (conductor-filled) height h_{su} is obtained from the following:

$$A_{su} \approx \frac{h_{su}}{2} \left(W_s + \frac{\pi(D_{is} + 2h_{su} + 2h_{ss})}{6 p_1 q_1} - (\tau_s - W_s) \right) \quad (10.105)$$

$$D_{os} = D_{is} + 2(h_{su} + h_{ss}) + 2h_{ys} \quad (10.106)$$

Note that if it is required to increase the machine inductance, the winding may be placed only in the upper part of the slot (Figure 10.46). In this latter case, Equation 10.95 has to be modified to produce the correct value of the slot conductor-filled slot height $h'_{su} < h_{su}$.

As the machine has PM poles on the rotor, it is characterized by a single synchronous inductance L_m :

$$L_m = \frac{6\mu_0 (W_a K_{W1})^2 \cdot \tau \cdot L(1 + K_{fringe})}{\pi^2 p_1 g_{Mt}} \quad (10.107)$$

The leakage inductance L_{sl} [25] has to be added to complete the synchronous inductance expression $L_s = L_m + L_{sl}$. K_{fringe} accounts for the flux fringing in the airgap in the radial plane and in the axial plane. The smaller the τ/g_{Mt} and λ ratios are, the larger the k_{fringe} . FEM flux distribution calculations are required to calculate k_{fringe} with precision.

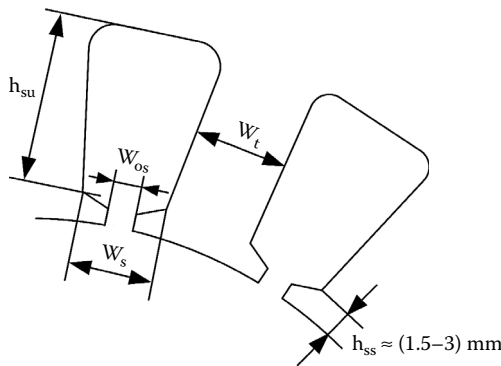


FIGURE 10.46 Stator slotting.

10.12.3 The Losses

The loss components in PMSG are as follows:

- Fundamental stator core losses
- Fundamental stator copper losses
- Rotor electric losses
- Mechanical losses
- Additional stator copper losses
- Additional stator core losses

Due to the large fundamental frequency and the contents in time harmonics of stator currents, the additional stator core and copper losses are notable. Also, the rotor electric losses have to be reduced to avoid PM overheating due to eddy currents in the PMs caused by stator space field harmonics and time current harmonics. The thin copper shield is a solution to the problem that results not only in PM thermal protection, but also in lower total rotor electric losses [26]. As a bonus, the rotor yoke may be made of solid iron without inflicting notable additional yoke losses through eddy currents.

A complete machine model to account for the rotor electric losses and additional stator core losses may be analytically developed.

The fundamental stator core and copper losses are straightforward [25], but the additional winding losses caused by the skin effect are of special interest. When the fundamental frequency rises above 500 to 600 Hz, full transposition of elementary paralleled conductors or the Litz wire are the ideal solutions to reducing the skin effect copper losses to negligible values. However, such a solution is costly.

It may as well serve the purpose, up to 2.2 kHz fundamental frequency and up to 200 kW, to parallel elementary conductors and use only 180° transposition (twisting turns only at one end-connection side).

A 20 to 30% increase in fundamental copper losses due to skin effect (“self-inflicted” and through circulating eddy currents) may be obtained this way with a two-layer winding with chorded coils.

For powers above 200 kW and fundamental frequency above 0.8 to 1.0 kHz, refined transposition, as done in large SGs, may be necessary to secure less than 33% of skin-effect copper losses in excess to fundamental copper losses.

Mechanical losses are also notable at high speeds, and they require special treatment for an in-depth investigation. They depend on the rotor diameter, length, peripheral speed, mechanical airgap, and bearing type (mechanical, with air or active magnetic).

A rough approximation for p_{mec} is as follows:

$$p_{mec} = C_0 \cdot D_r \cdot l_{rotor} (\pi \cdot D_r \cdot n)^2; \quad D_r = D_{is} - 2g \quad (10.108)$$

The constant $C_0 \approx 5 \text{ W s}^2/\text{m}^4$ for air-cooled generators. D_r is the rotor diameter, and n is the rotor speed (rpsec). For additional information on super-high-speed PM synchronous machines, see the literature [27–33].

Note that the PM generator/motor for flywheel energy storage is a typical case of a super-high-speed such machine [34].

10.13 Super-High-Speed PM Generators: Power Electronics Control Issues

The power electronics control of super-high-speed PM generators depends on the following:

- Operation modes: generating only or self-starting (motoring)
- Constant power operation speed range at constant output voltage and frequency
- Power level and fundamental rated frequency

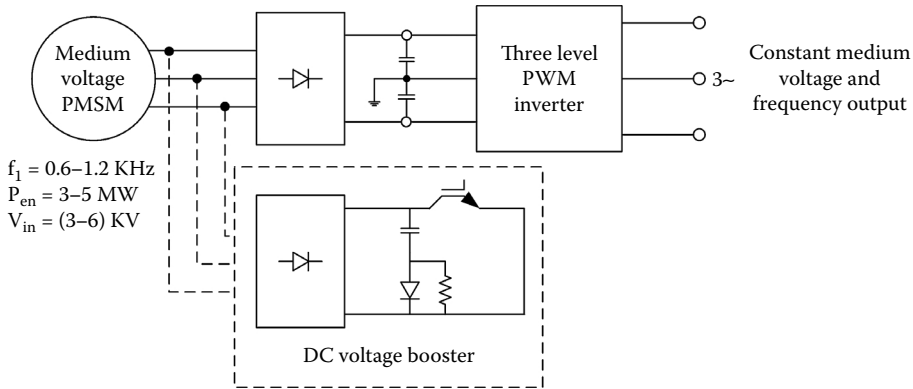


FIGURE 10.47 A 3 to 5 MW medium-voltage super-high-speed permanent magnet synchronous generator (PMSG) ($f_1 \approx 0.6$ to 1.2 kHz) with direct current (DC) voltage booster and three-level pulse-width modulated (PWM) inverter.

Let us discriminate between machines below 2 to 3 MW where standard six-leg PWM low-voltage source IGBT converters (up to 690 V line voltage [RMS]) are already produced, and powers up to 5 to 6 MW at 15,000 to 18,000 rpm, envisaged for the near future where, most probably, medium voltage (2 to 5 kV) multilevel inverters will be required. In general, in the MW power range, generator-only operation is more likely.

Then, with the fundamental frequency in the 0.6 to 1.2 kHz range, above 2 MW, in a medium-voltage PMSG design (3.0 to 6 kV line voltage RMS), a general control system configuration comprises the following (Figure 10.47):

- A front machine end diode rectifier without or with a boost converter
- A multilevel voltage-source inverter for constant frequency and voltage output

The DC voltage booster is mandatory if the speed range for constant power is larger than 110 to 120%. In Figure 10.47, the DC voltage booster makes use of the generator inductances for energy storage through a fast recovery additional diode rectifier short-circuited in a controlled manner through an oversized IGBT power switch at a frequency of at least 8 to 10 kHz, as the fundamental frequency may reach 1.2 kHz. This way, the voltage in the DC link is kept constant over the entire speed range of interest and for all load (and overload) conditions.

Three-level PWM inverters represent a strong emerging technology with already notable industrial success. There are quite a few basic categories of multilevel PWM inverters [35]:

- Cascaded multicell inverters
- Flying capacitor inverters
- Diode-clamp (neutral point clamp) inverters

All of these categories [35] have merits and demerits, but the diode-clamp multilevel PWM converters seem to be the first-choice solution for the case in point because of more industrial experience with MW drives at 2.3, 3.3, 4.16, and 6 kV.

A typical three-level diode-clamp inverter leg is shown in Figure 10.48 with only AC terminal *a* shown. Because the inverter is now used as a power source, careful output filtering of the output is required. Such output filters are different, in general, for power grid and stand-alone operation.

Vector control or direct active and reactive power control may be applied here as done for two-level standard PWM converters [36–39].

For powers below 1.5 MW, low-voltage (690 V line RMS voltage) PMSGs of super-high speeds may be built.

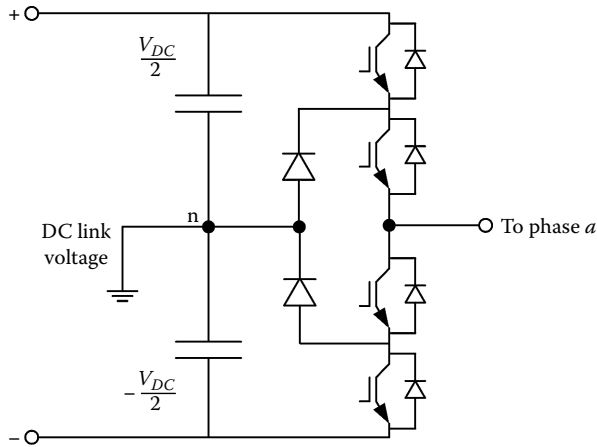


FIGURE 10.48 Three-level diode-clamped pulse-width modulated (PWM) inverter.

A two-level (standard) PWM voltage source IGBT converter may be used for this case. For four-quadrant operation (self-starting), the use of two back-to-back PWM voltage source converters is the first choice (Figure 10.49).

The control of the cascaded AC–AC converter is similar to the case of the cage rotor induction generator.

In essence, for power grid operation, the grid-side converter may be vector controlled to keep the DC voltage rather constant (along axis q) and to produce a certain value of reactive power (along axis d).

The control of the machine-side converter for this situation may be implemented as vector control or as direct torque control. For a wind turbine, a torque/speed reference produces, for each value of speed, a certain optimum reference torque. The machine stator flux reference may be set constant with speed, and finally, direct torque and flux control may be performed (Figure 10.50a through Figure 10.50c).

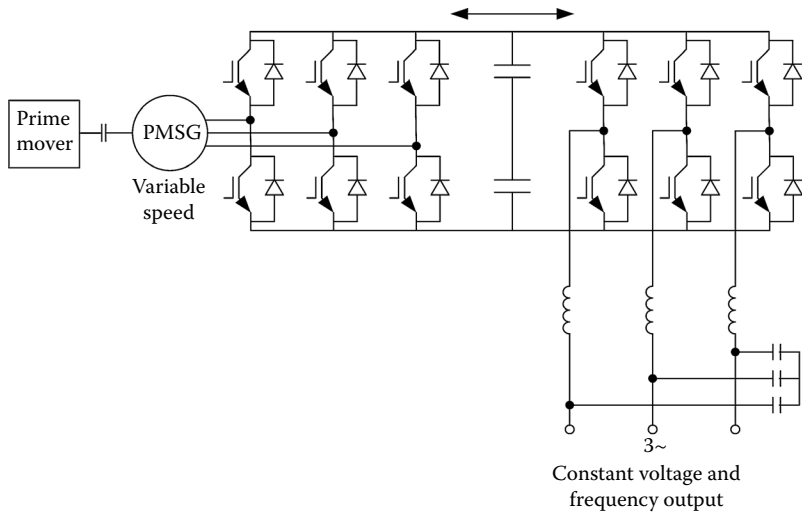


FIGURE 10.49 Four-quadrant cascaded alternating current (AC)–AC converter.

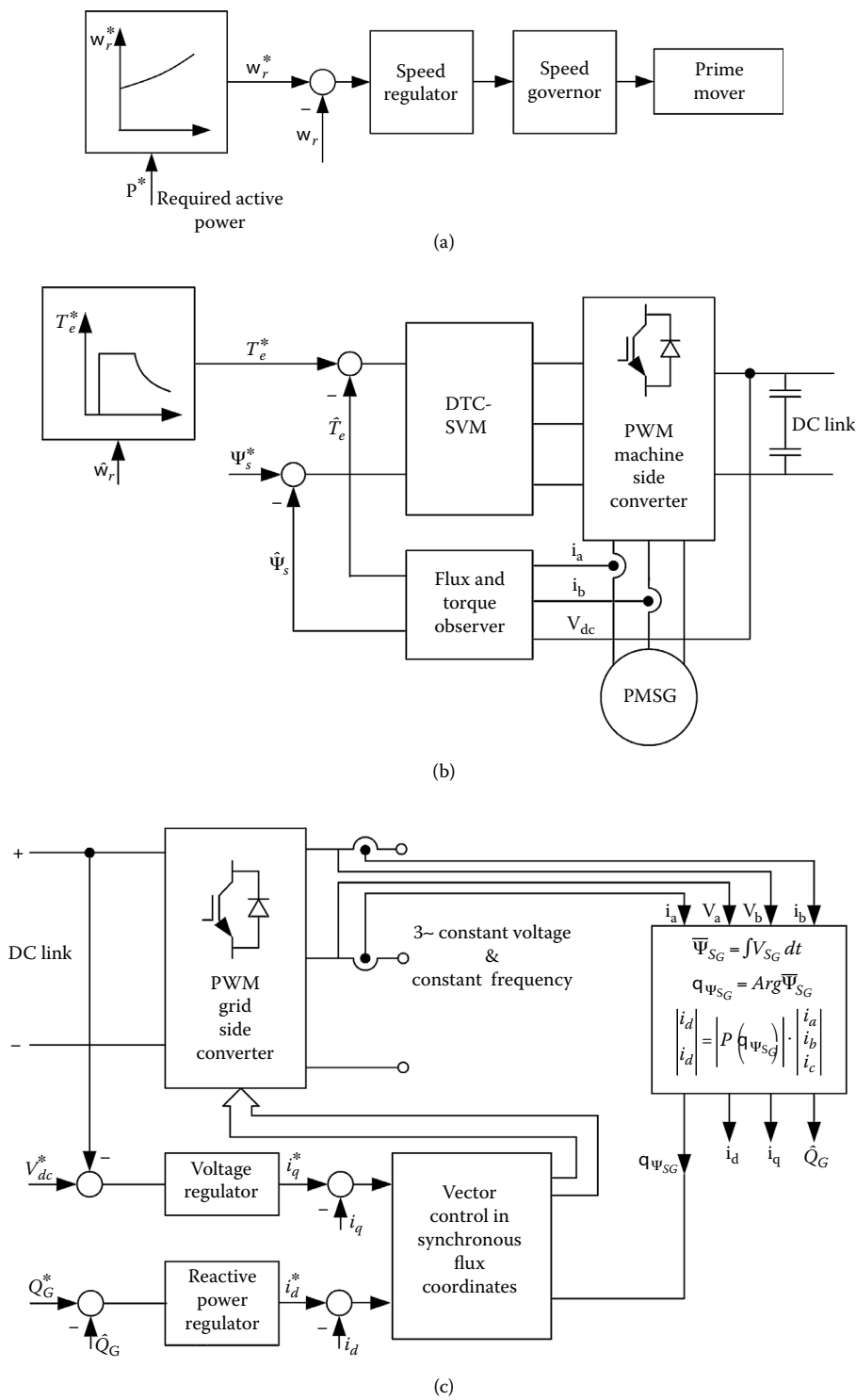


FIGURE 10.50 Generic four-quadrant cascaded alternating current (AC) converter-fed permanent magnet synchronous generator (PMSG) at the power grid: (a) speed control by prime mover, (b) direct torque and flux control of machine-side converter, and (c) vector control of grid-side converter.

The speed of the prime mover should be close-loop controlled through the speed governor, having the optimum speed required power, known off-line (Figure 10.50a).

As a fundamental frequency goes up to 1 to 2.5 kHz, the question arises, again, as to whether trapezoidal current control [39] in the generator is not more practical in reducing the commutation losses in the machine-side converter, especially if a DC–DC converter is added to control the DC link voltage for variable speed when the PM generator works alone or is connected to a weak local power system.

10.14 Design of a 42 V_{dc} Battery-Controlled-Output PMSG System

In the previous section, dedicated to super-high-speed PMSGs, a few design (sizing) issues were treated in some detail. In general, PMSG sizing has to be performed with full consideration of converter costs and losses as also calculated for the critical duty cycles of motoring and generating vs. speed.

For example, the design of a PM reluctance synchronous generator/motor for a mild hybrid vehicle with a 42 V_{dc} battery and a 6:1 motor/generator constant power speed range (from 1000 rpm to 6000 rpm) [40, 41] is quite different from the design of an automotive PMSG with a 6:1 constant power speed range in generating only.

The increase in electric power demand on board of pure internal combustion engine (ICE) vehicles today seems to establish the PMSG with single IGBT control and 42 V_{dc} battery for powers above 1.5 kW as a potential replacement for the claw-pole-rotor generator. But, for 50 million cars per year, the production potential seems staggering.

Fast power control at higher efficiency is the main merit of the automotive PMSG.

In view of the above, in what follows, we will treat in some detail the design of the 42 V_{dc} battery controlled output automotive PMSG.

The design process generally includes the following:

- Design initial data
- Topologies of interest
- Electromagnetic design
- Thermal verification
- Design output data

10.14.1 Design Initial Data

For the case in point, only the rated power P_{en} and the battery voltage V_{dc} are defined from the start.

A 10:1 speed range is typical, but it may be required to deliver only 50% of P_{en} at minimum speed n_{min} and 100% P_{en} above $2n_{min}$ and up to n_{max} [40]. A more stringent requirement corresponds to 100% P_{en} delivery at n_{min} [40]. It is well understood that in this latter case, the machine has to deliver twice as much torque in comparison to that in the first case.

As the torque “dictates” the machine size, together with losses, and, as volume is severely constrained, the 50% P_{en} at n_{min} requirement is tempting and corresponds to existing practice with claw-pole-rotor alternators on board vehicles.

10.14.2 The Minimum Speed: n_{min}

The second important initial data include the speed range.

The ICE idle speed is around 800 rpm. As the larger the PMSG speeds, the lower its volume; use of a belt transmission to raise the PMSG speed seems to be a practical solution. But, what should be the belt transmission ratio k_{belt} ?

In Reference 40, to determine n_{min} of PMSG, the total losses in the generator are made functions of speed n and are then minimized to find the optimum n_{min} .

In essence, the PMSG losses are divided into the following:

- p_{trans} : by vehicle deceleration (mechanical)
- p_{rota} : by generator deceleration (mechanical)
- p_d : in the rectifier
- p_{iron} : iron losses
- p_{co} : copper losses
- p_{fan} : fan losses

The p_{trans} and p_{rota} are to be calculated as corresponding to the standard town driving cycle.

In general, the torque of the generator is proportional to machine volume (r^3), and thus, at minimum generator speed n_{min} ,

$$P \sim r^3 n_{min} \quad (10.109)$$

Consequently, from Equation 10.109,

$$r \approx n_{min}^{-1/3} \quad (10.110)$$

So, the generator mass m is

$$m \sim n_{min}^{-1} \quad (10.111)$$

The translational energy of the generator is reduced when the vehicle decelerates from U_v to U_{v-1} :

$$p_{tran} = \sum_v \frac{m}{2} (U_v^2 - U_{v-1}^2) \quad (10.112)$$

Also, when the generator decelerates from n_v to n_{v-1} , the losses related to this rotation deceleration are as follows:

$$p_{rota} = 2\pi^2 J \sum_v (n_v^2 - n_{v-1}^2) \quad (10.113)$$

with J equal to the rotor inertia.

The whole town driving cycle has to be considered to find average values for p_{trans} and p_{rota} :

$$p_{trans} \sim m \sim n_{min}^{-1} \quad (10.114)$$

$$p_{rota} \sim J \sim r^5 \sim n_{min}^{-5/3} \quad (10.115)$$

The iron losses in the generator have two components: hysteresis and eddy current losses. As hysteresis losses are proportional to speed (frequency), and the mass is proportional to n_{min}^{-1} , it follows that the former are independent of speed. The eddy current losses, however, are proportional to mass and speed squared, and thus,

$$p_{iron} \approx m n_{min}^2 \sim n_{min} \quad (10.116)$$

The copper losses depend on the skin effect if the frequency goes higher than 500 to 600 Hz, but they may be assumed to be proportional to the radius of the rotor:

$$p_{co} \sim r \sim n_{min}^{-1/3} \quad (10.117)$$

Finally, the fan losses are considered to be proportional to r^4 :

$$p_{fan} \sim r^4 \sim n_{min}^{4/3} \quad (10.118)$$

If the coefficients in the above formulas are known, the total losses in generator $\sum p$ may be minimized with respect to n_{min} :

$$\sum p = C_{tran} n_{min}^{-1} + C_{rota} n_{min}^{1/3} + C_{iron} n_{min} + C_{co} n_{min}^{-1/3} + C_{fan} n_{min}^{4/3} \quad (10.119)$$

As the information to define the constants in Equation 10.119 is not very reliable, it is safe to choose a few values of n_{min} and compare the final design results. However, it seems that $n_{min} = 1800$ to 3000 rpm is a realistic range. The belt transmission ratio thus varies as $k_{belt} = (1800/800) \div (3000/800) = 2.25$ to 3.75. With the increase in material performance and power electronics switching frequency, the tendency should be toward $n_{min} = 3000$ rpm with $n_{max} = 30,000$ rpm.

10.14.3 The Number of Poles: $2p_1$

Increasing the number of poles of the PMSG tends to reduce the stator coil end turns and the stator yoke height (and weight) but leads to higher fundamental frequency f_1 :

$$f_1 = p_1 \cdot n \quad (10.120)$$

The higher frequency generally leads to an increase in core loss despite a reduction of yoke height. Therefore, the copper losses tend to decrease, but the core losses tend to increase with frequency for given speed.

The fast recovery diodes used in the diode rectifier, most likely to be used in the PMSG system, together with a single IGBT chopper, allow for a maximum fundamental frequency in the range of $f_{1max} = 1.5$ to 2.5 kHz.

For $n_{max} = 30,000$ rpm and $2p_1 = 10$ poles, the frequency $f_{1max} = 5 \cdot 30 \times 10^3 / 60 = 2.5$ kHz.

As, currently, super-high-speed AC drives with fundamental frequency in this range have been built up to 100 kW and more, there is enough credibility to pursue this way of thinking for car alternators.

Concerning the stator core laminations used for the scope, it was found that even 0.36 mm thick 3.1% silicon laminations with special thermal treatment may cause only 30 W/kg core losses at 1 T and 800 Hz. Reducing the silicon lamination thickness to around 0.1 to 0.12 mm, while maintaining a 0.8 to 0.85 filling factor with a superthin insulation coating, can keep the specific iron losses to about 30 W/kg, at 1 T and 2.5 kHz fundamental frequency. Finally, it is worth considering fabricating the stator core from soft magnetic composites (SMCs), with permeability that has reached $\mu_{rel} = 500$ P.U. [42] at 1 T. The hysteresis losses in such materials are independent of frequency, while the eddy current losses increase linearly with frequency. The SMC losses become smaller than those for silicon laminations above 500 to 700 Hz.

Though the relative permeability is still notably smaller than that for silicon steel laminations, the use of thick surface PM poles on the rotor of the PMSG leads to acceptable airgap flux density reduction as long as the flux path length in iron is not too long. Consequently, a large number of poles favor the application of SMC.

The possibility of compressing the stator core with windings in slots through electromagnetic force [43] in case of SMC, may lead to further increases in slot filling factor and to geometry reduction.

The above rationale leads to the conclusion that the number of poles $2p_1$ may be as large as ten for PMSGs dedicated to automotive applications with fundamental maximum frequencies up to 2.5 kHz.

10.14.4 The Rotor Configuration

The rotor configuration may be of cylindrical shape (with radial airgap) or of disk shape (with axial airgap). There are pros and cons for both geometries [40, 41]. However, it is considered here that the cylindrical rotor is more rugged, and the stator is easier to fabricate. Further, the surface PM rotor configuration was proven to produce higher torque per PM volume. Also, it is the configuration that allows for a solid rotor yoke (back iron), provided a thin copper screen is placed over the PMs. In Section 10.12.1, this configuration was introduced (Figure 10.45).

The surface PM poles on the rotor are to be made of parallelepipedic PMs, tangentially and axially, in order to reduce the costs of the PMs. The copper thin screen over the PMs is meant to reduce the PM and back iron eddy current losses due to stator mmf and slot opening space harmonics and due to stator current time harmonics at the cost of some, but reduced, losses in the copper screen, which, at high frequencies, behaves like a predominantly reactive circuit.

The retaining carbon fiber epoxy resin ring is used to protect the PMs and the copper shield from centrifugal forces, as the rotor peripheral speed reaches 150 to 200 m/sec for the case in point. The solid back iron of the rotor provides for a rugged rotor structure (Figure 10.51).

10.14.5 The Stator Winding Type

The stator winding may be distributed, in general, for $2p_1 = 4, 6$ and $q = 2$ slots/pole/phase, when chorded coils with $Y/\tau = 5/6$ are used. To cut the costs of winding and copper end-turn losses, nonoverlapping coils may be used.

For $2p_1 = 8$, the number of stator slots N_s should be 9 or even 12, while for 10 poles, $N_s = 12$. In these cases, the equivalent winding factor (as shown in Table 10.1) is $k_{w_1} = 0.945$. The cogging torque is also low, as the smallest common multiplier (SCM) of N_s and $2p_1$ is rather high: 72 and, respectively, 60.

Two coils are inserted in a slot, and thus, the end-turn axial extension and, consequently, the frame axial length are further reduced (Figure 10.52a and Figure 10.52b).

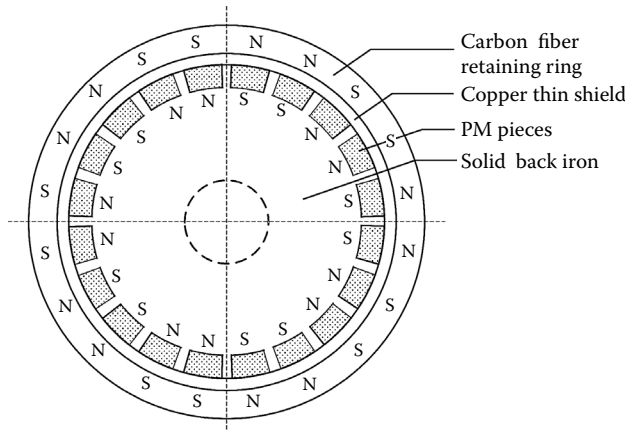


FIGURE 10.51 Ten-pole permanent magnet (PM) rotor.

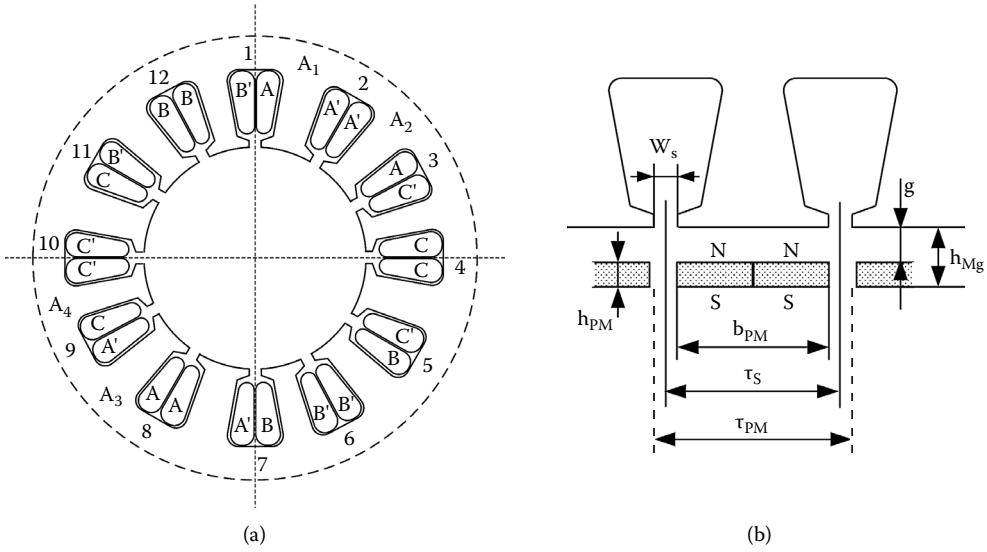


FIGURE 10.52 (a) Ten-pole 12-slots two-layer winding and (b) permanent magnet (PM) slot geometry.

For the 12/10 ($N_s/2p_1$) combination (Figure 10.52),

$$\frac{\tau_s}{\tau_{PM}} = \frac{10}{12} \quad (10.121)$$

To further reduce cogging torque, in general, the PM span b_{PM} is taken around

$$b_{PM} \approx \tau_s - W_{os} \quad (10.122)$$

As the design of a PMSG for $2p_1 = 4$ and distributed winding is presented in detail in Reference 42, we will concentrate more on the case of nonoverlapping windings.

It was noticed that a sinusoidal emf may also be produced with this winding by carefully calibrating W_{os} to magnetic airgap h_{Mg} and PM span b_{PM} . In general, the slot opening W_{os} is smaller than the total magnetic airgap h_{Mg} , in order to reduce the slot opening reinforcement of stator mmf harmonics flux density as felt on the copper shield upper surface. This way, the copper shield losses are reduced.

Also, a smaller slot opening means a relatively larger slot leakage inductance, which influences the maximum torque that can be produced by the PMSG, due to an enlarged synchronous inductance L_s .

10.14.6 Winding Tapping

For 5:1 constant power speed range, winding tapping may not be ruled out as a means to reduce the emf at maximum speed and thus allow for a lower voltage rating of the single IGBT chopper that will control the output DC current received by the battery and load.

It is a 5:1 speed range from $2n_{min}$ to n_{max} , while at n_{min} , only half the power is delivered. The simplest way to achieve the winding tapping is by dividing it into two halves. The two halves work in series up to about $n_{max}/2$. After that, only one half remains active. To achieve this goal, two diode rectifier bridges and an electromagnetic power switch are required (Figure 10.53).

The diode rectifier D_{low} works below $n_{max}/2$ with power switching (PS) closed, while D_{high} works at high speeds (above $n_{max}/2$) with PS open.

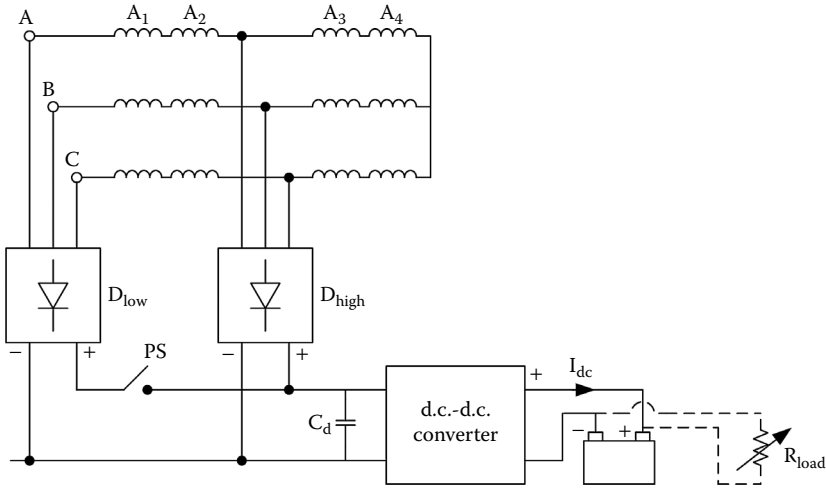


FIGURE 10.53 Winding tapping with two diode rectifiers and an electromagnetic power switch (PS).

For a 12/10 combination winding, one of the two neighboring coils pertaining to each phase is grouped in one of the two winding halves A_1 – A_2 and A_3 – A_4 in Figure 10.51. In this way, a uniform mmf reaction field is secured even when only half of the winding is active above $n_{max}/2$.

10.14.7 The PMSG Current Waveform

Depending on the machine inductance L_s , DC link capacitor C_d , and the fundamental frequency (speed), the generator phase current may be continuous (almost sinusoidal) or discontinuous (trapezoidal). The DC–DC converter controls the current in the generator, according to load needs.

In Reference 40, it is inferred that the generator current is discontinuous, though only test waveforms at 3000 rpm are exhibited to prove it. Moreover, the distributed windings in Reference 40 are slotless, and thus, the machine inductance L_s is smaller than usual.

The winding tapping increases the phase current above $N_{max}/2$ when it otherwise tends to be smaller, because the emf increases with speed.

10.14.8 The Diode Rectifier Imposes almost Unity Power Factor

Considering that the emfs per phase are sinusoidal and $L_s = L_d = L_q$, the d – q model (Equation 10.76 and Equation 10.77) applies, and for almost zero Q_1 (reactive power), Equation 10.79 becomes

$$L_s (i_d^2 + i_q^2) - \Psi_{PM} i_d = 0 \quad (10.123)$$

The maximum torque conditions (Equation 10.81 and Equation 10.82) apply to give the following:

$$i_{dk} = \Psi_{PM} / 2L_s = i_{qk} = i_{sk} / \sqrt{2} \quad (10.124)$$

$$T_{ek} = \frac{3}{2} p_1 \frac{\Psi_{PM}^2}{2L_s}$$

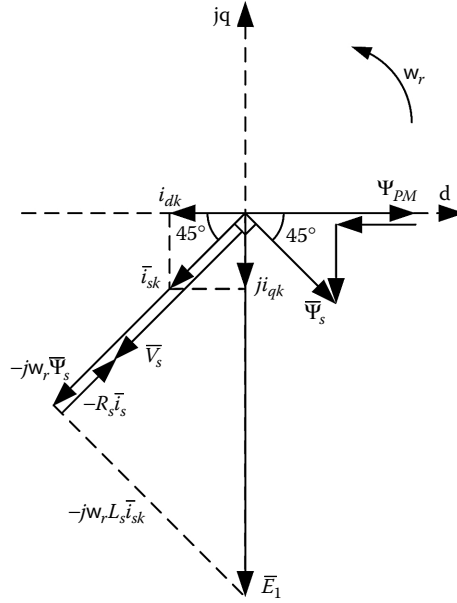


FIGURE 10.54 Vector diagram at maximum torque and unity power factor.

10.14.9 Peak Torque-Based Sizing

At n_{min} , corresponding to the ICE idle speed (n_{min}/k_{belt}) with half the rated power delivered, and with full power at $2n_{min}$, it seems that the design for torque has to be performed for n_{min} and $P_{en}/2$:

$$T_{ek} = \frac{(P_{en}/2)}{\eta_{n_{min}} 2\pi n_{min}} \quad (10.125)$$

The maximum torque condition (Equation 10.124) may be used for sizing the machine. An assigned value of efficiency at n_{min} (0.7 or so) and T_{ek} has to be put forward only to be corrected later in design.

For sinusoidal current, the voltage equation (Equation 10.76) yields the vector diagram in Figure 10.54:

$$\begin{aligned} \bar{I}_s R_s + \bar{V}_s &= -j2\pi p n_{min} \bar{\Psi}_s = -j\omega_{rmin} \bar{\Psi}_s \\ \bar{\Psi}_s &= \Psi_{PM} + L_s i_{dk} + jL_s i_{qk}; \quad i_{dk} = i_{qk} = \Psi_{PM}/2L_s \end{aligned} \quad (10.126)$$

As known, V_s is equal to $V_{lph}\sqrt{2}$, where V_{lph} is the phase RMS voltage.

10.14.10 Generator to DC Voltage Relationships

The relationship between V_s and the diode rectifier voltage V_{dc} is as follows:

$$V_{dc} \approx \frac{3\sqrt{6}}{\pi} E_{lph} - \Delta V_{diode} - \frac{3}{\pi} L_c \omega_{lmin} I_{dc} \quad (10.127)$$

ΔV_{diode} is the voltage drop along the diodes in series between two PMSG phases. L_c is the commutation inductance of PMSG. L_c corresponds to the subtransient inductance $L_s'' \approx L_s$. However, it should be maintained that the voltage drop along L_c in Equation 10.127 is valid with discontinuous current and E_{1ph} :

$$E_{1ph} \sqrt{2} = \omega_r \Psi_{PM} \quad (10.128)$$

For sinusoidal current, the commutation voltage drop in the diode rectifier is “replaced” by the voltage drop along the machine reactance $\omega_1 L_s$, and then,

$$V_{dc} \approx \frac{3\sqrt{6}}{\pi} V_{1ph} - \Delta V_{diode} \quad (10.129)$$

$$V_{1ph} = -R_s \frac{I_{sk}}{\sqrt{2}} + \sqrt{\left(\frac{\omega_{rmin} \cdot \Psi_{PM}}{\sqrt{2}} \right)^2 - \omega_{rmin}^2 L_s^2 I_{sk}^2 / 2} \quad (10.130)$$

$$I_{sk} = i_{dk} \sqrt{2} = \frac{\Psi_{PM}}{2L_s} \sqrt{2}$$

Equation 10.127 through Equation 10.130, together with the power balance, are crucial for the design at ω_{rmin} and T_{ek} :

$$\frac{P_{em}}{2} = V_{dc} I_{dc} = 3V_{1ph} \cdot \frac{I_{sk}}{\sqrt{2}} - \Delta V_{diode} I_{dc} \quad (10.131)$$

10.14.11 The Ψ_{PM} , L_s , R_s Expressions

The stator PM flux Ψ_{PM} is:

$$\Psi_{PM} = \phi_{Pmax} \times W_A K_{W1} \quad (10.132)$$

The maximum PM flux per phase pole (tooth) is as follows:

$$\phi_{Pmax} = B_{gPM}^a \cdot \tau_s \cdot l_{stack} \cdot K_{Fc} \quad (10.133)$$

B_{gPM}^a is the average airgap PM flux density below the stator slot pitch (for the nonoverlapping coil winding) when the PM pole has its axis along that stator tooth (pole) axis. Also, l_{stack} is the magnetic length of the stator stack; W_a is the number of turns in series per phase (or per current path). For the 12/10 ($N_s/2p_1$) combination, $W_1 = 4n_c$ if all the coils per phase are connected in series.

Again, the equivalent winding factor for the 12/10 ($N_s/2p_1$) combination is 0.945 according to [Table 10.1](#).

The cyclic machine inductance contains the self-inductance and a contribution of the mutual inductance between adjacent phases.

Because for the 12/10 ($N_s/2p_1$) combination half of the flux is closed through adjacent phases, the cyclic inductance L_s becomes as follows:

$$L_s \approx L_{sl} + \frac{3}{2} L_{ss} \quad (10.134)$$

L_{ss} is the airgap self-inductance per phase:

$$L_{ss} = \frac{\mu_0 (W_a K_{W1})^2 \tau_s l_{stack}}{\left(\frac{N_s}{3}\right) h_{Mg} K_c (1 + K_s)} (1 + K_{fringe}) \quad (10.135)$$

where

h_{Mg} is the total magnetic airgap

K_c is the Carter coefficient

K_s is the magnetic core saturation coefficient

The leakage inductance L_{sl} is calculated with a standard expression and essentially contains slot leakage and end-turn leakage:

$$L_{sl} = c_{sl} \cdot W_a^2 \quad (10.136)$$

L_{ss} tends to be larger than for the non-overlapping windings with equivalent data by as much as 20 to 30% or more, mainly due to additional differential leakage inductance additions. In view of the large total magnetic airgap, the magnetic saturation coefficient k_s is likely to be smaller than 0.25.

The Carter coefficient accounts for slot openings and is likely to be close to unity, as $W_{os} < h_{Mg}$ ($1 < K_c < 1.1$).

As the slot openings are rather small and the total magnetic airgap is large, L_{sl} is a good part (up to, maybe, 66%, in some cases) of L_{ss} .

The stator phase resistance R_s is as follows:

$$R_s = \frac{N_s}{3} \rho_{c0} \cdot \frac{l_{coil}}{A_{cooper}} (1 + K_{skin}); \quad l_{coil} \approx 2l_{stack} + \tau_s (1 + \pi/2) \quad (10.137)$$

The cross-section of the conductor from which the stator turns are made is as follows:

$$A_{cooper} = \frac{I_{sk} / \sqrt{2}}{a j_{cok}} \quad (10.138)$$

where j_{cok} is the peak current density adopted for peak torque ($P_{en}/2$ at n_{min}).

Given the large maximum frequency, elementary conductors in parallel should be used to reduce skin effect. Moreover, at least 180° transportation (twisting of wires only in one coil end-connection zone) is required to reduce the circulating currents due to proximity skin subeffect in paralleled conductors. In general, k_{skin} has to be brought to below 0.33, as done for large SGs that face a similar problem at low frequency due to very large conductor cross-section.

The design approach in the previous paragraph corroborated with the methodology in this paragraph should provide a solid basis for a practical design. We mention here only that peak torque T_{ek} expressions (Equation 10.124 and Equation 10.125) are the basis for stator bore diameter and voltage and power expressions (Equation 10.128 and Equation 10.129) and are to be used to find the number of turns per phase (coil).

We alluded to the design methodology in Reference 41 and stipulated above a few new design insights for the envisaged application. We now stop here, leaving the interested reader the freedom to further our pursuit the way he or she feels fit.

10.15 Methods for Testing PMSGs

PMSGs represent a rather novel technology, and thus, special standards for their testing are not yet available. They may be, however, assimilated with SGs with standards that are available (Institute of Electrical and Electronics Engineers [IEEE] standard 115/1995). Even for SGs, the standards refer to constant speed operation.

On the other hand, methods for testing PM synchronous motors were proposed over two decades ago [45, 46].

The methods of testing PMSGs can be classified by the operation modes:

- Standstill modes
- No-load and short-circuit tests
- Load tests

or by the scope:

- For loss and efficiency assessment
- For parameters estimation

Standstill tests are used only for parameter estimation, while the others may be used either for parameter or for loss (efficiency) assessment.

The types of tests required for parameter (L_d , L_q) estimation depend on the following:

- The presence (or absence) of damper (copper shield) on the rotor
- The surface PM or IPM rotor pole configuration, which helps to determine if magnetic saturation is or is not important

We will proceed with standstill tests and then continue with no-load and short-circuit tests to end with load tests and emphasize which parameters or losses can be identified there.

10.15.1 Standstill Tests

The DC current decay and frequency response in axial d and q standstill tests were standardized and are described in the IEEE standard 115/1995, dedicated to general SG.

The DC current decay tests may produce the magnetization curves along axis d and q by integrating the voltage drop along stator resistance (Figure 10.55a through Figure 10.55c and Figure 10.56a through Figure 10.56c):

$$L_d(i_d)i_{d0} = \int_0^\infty R_s i_d dt + \frac{2}{3} \int_0^\infty V_d dt; \quad i_d(t) = i_a(t) \quad (10.139)$$

$$\Psi_d(i_{d0}) = \Psi_{PM} + L_d(i_{d0})i_{d0} \quad (10.140)$$

The rotor is placed in axis d by itself when the stator is DC current fed with the connection of phases as in Figure 10.55a. The rotor is then stalled into the d axis position.

The tests in axis q are made without moving the rotor from the initial position by changing the connection of phases, as in Figure 10.56:

$$\Psi_q(i_{q0}) = L_q(i_{q0})i_{q0} = \frac{\sqrt{3}}{2} \left[\int_0^\infty R_s i_b dt + \frac{1}{2} \int_0^\infty V_d dt \right] \quad (10.141)$$

The experiment has to be done for a few initial values of DC current — positive and negative in axis d (along PMs) and only positive in axis q .

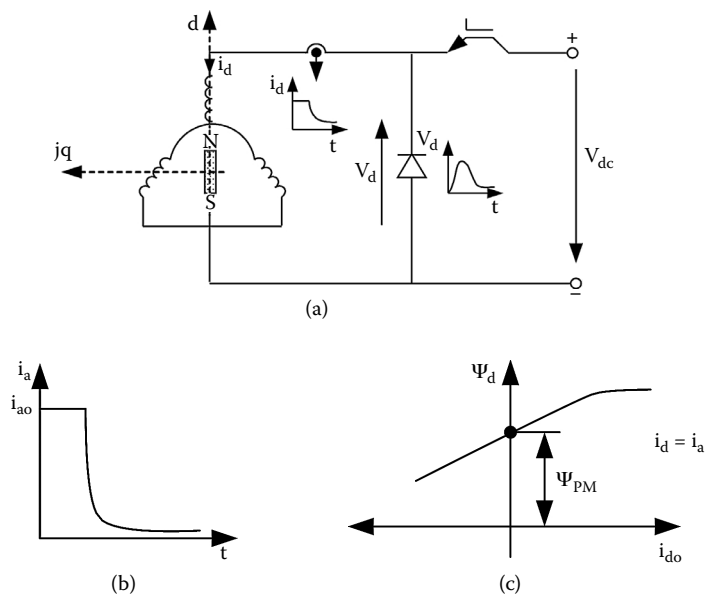


FIGURE 10.55 Standstill tests in axis d : (a) the scheme, (b) the current decay, and (c) d axis magnetization curve.

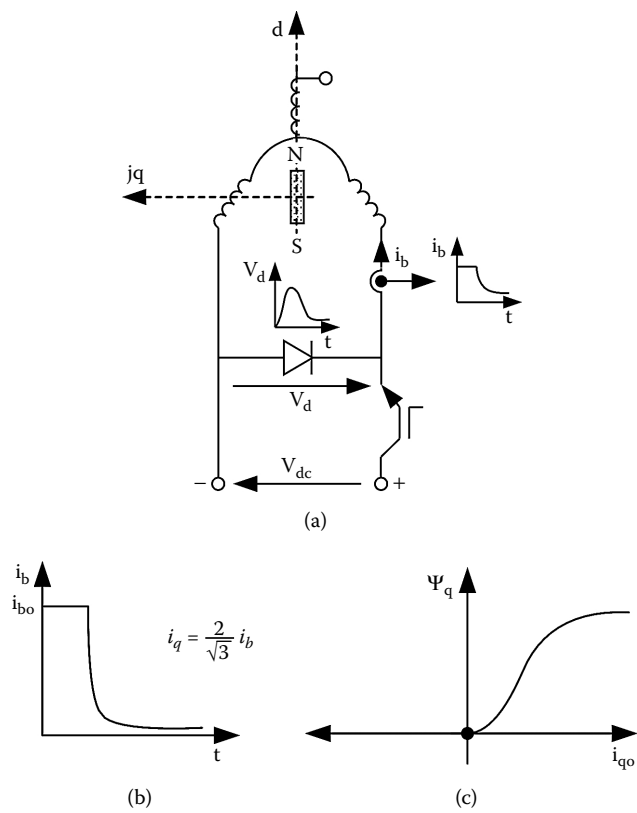


FIGURE 10.56 Standstill tests in axis q : (a) the scheme, (b) the current decay, and (c) q axis magnetization curve.

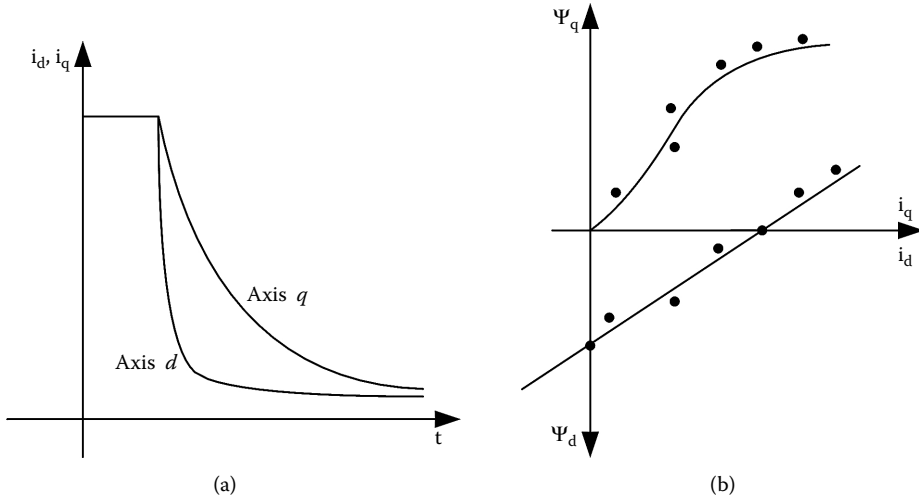


FIGURE 10.57 (a) Test results in direct current (DC) decay tests in axes d and q (b) for a 140 Nm peak torque high-saliency interior permanent magnet (IPM) machine.

Note the following:

- The diode voltage drop has to be considered in the integral, as it may not be negligible, and thus, may introduce errors above 10 to 15%.
- The stator resistance R_s may vary from one test to another, and thus, it should be updated before each current decay test by measuring V_{dc} and i_a (i_b) and calculating $R_s = 2V_{dc}/3I_{a0}$; $R_s = V_{dc}/2I_{b0}$.
- The magnetic hysteresis may also influence the result. After each test, a few positive and negative ($\pm$) decreasing current pulses may be applied to cancel the hysteresis effects.
- The variable voltage (current) source with fast turnoff may be a standard voltage-source PWM inverter controlled with a variable modulation index to produce only the conduction of phases in Figure 10.55a and Figure 10.56a.
- In a surface PM pole rotor PMSG, $L_d = L_q$ and the d and q standstill DC current decay tests may be used to verify that this is the situation or to uncover an IPM rotor.
- Typical DC current decay test results, obtained according to the above methodology, are shown in Figure 10.57a and Figure 10.57b for a 140 Nm peak torque, for a high-saliency IPM rotor SG.

In axis d , the flux Ψ_d changes sign only, apparently demagnetizing the PMs. But this is not the case, as the leakage inductance L_{sl} is about equal to the magnetization axis inductance L_{dm} ($L_d = L_{dm} + L_{sl}$). Consequently, most negative flux in axis d is leakage flux.

FEM calculations with current components in both axes d and q led to unique magnetization curves in axes d and q as functions of total stator current i_s (Figure 10.58).

The cross-magnetization, visible mostly in IPM machines, is still mild and may be described by the unique magnetization curves concept [50]:

$$L_d(i_s) = \frac{(\Psi_d^* - \Psi_{PM})}{i_s}$$

$$i_s > 0 \text{ for } \Psi_d^* > \Psi_{PM}; \Psi_d = \Psi_{PM} + L_d(i_s) \cdot i_d \quad (10.142)$$

$$L_q(i_s) = \frac{\Psi_q^*}{i_s}; \Psi_q = L_q(i_s) \cdot i_q$$

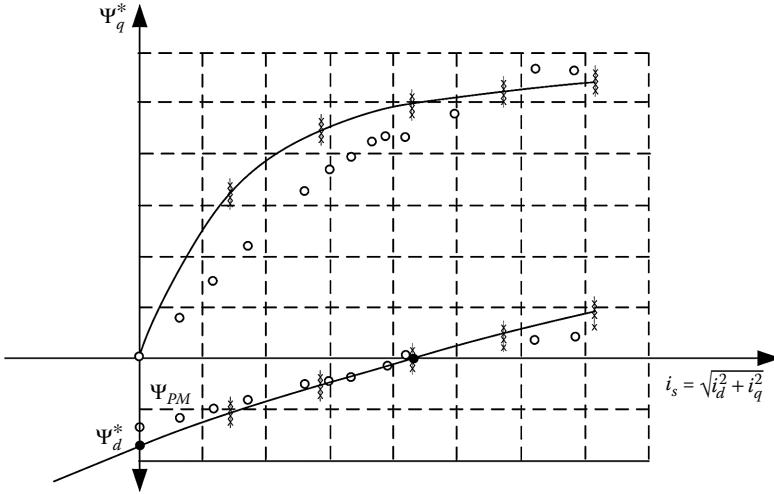


FIGURE 10.58 Finite element method (FEM)-extracted and experimental unique magnetization curves in axes d and q vs. total stator current i_s for an interior permanent magnet synchronous generator (IPM-SG). (Adapted from I. Boldea, L. Tutelea, and C.I. Pitic, *IEEE Trans.*, IA-40, 2, 2004, pp. 492–498.)

Equation 10.142 allows for the computation of d - q inductances after the unique magnetization curves are identified from FEM or from load tests, as explained later in this section.

In general, L_d (along the axis of PMs) may be considered constant, even in IPM machines, while L_q is a function of i_s (total stator current, when $i_d < i_q$) as obtained from standstill DC current decay tests.

For surface PM pole rotor machines, $L_d = L_q$, and both are constant or mildly variable as a function of i_s , rather than of i_d or i_q .

Let us note that, in reality, the above-described magnetization curve in axis d may not be drawn unless the PM flux linkage Ψ_{PM} is determined from a different test. At standstill, only if torque may be measured, the Ψ_{PM} may be determined from the following:

$$T_e = \frac{3}{2} p_1 (\Psi_d(i_s) i_q - \Psi_q(i_s) i_d) \quad (10.143)$$

If the rotor position θ_{er} and stator DC current i_a , with the phase connection as in [Figure 10.55a](#), are known,

$$i_d = i_s \cos \theta_{er}; \quad i_q = i_s \sin \theta_{er} \quad (10.144)$$

From the two unique magnetization curves, for given i_d , i_q , and measured torque T_e , $\Psi_d(i_s)$ is obtained with a few values of i_s and θ_{er} .

To measure the torque at standstill, either a torquemeter is used or a PM DC motor is loading the PMSG to maintain zero speed. The DC motor current i_{dcM} is proportional to the torque, and calibration of i_{dcM} to torque for the PM DC motor is straightforward: $T_{e_{dcM}} = k_T \cdot i_{dcM}$.

The frequency response tests at a single frequency may not be enough to determine the transient inductances L'_d , L'_q of the PMSG. The PMs represent a weak damper winding, while the copper rotor shield is also a mild one. We have to operate with frequencies in the range of those caused by the space mmf slot-opening harmonics and the current time harmonics corresponding to the entire speed range of the machine. For super-high-speed PMSGs, such frequencies are in the range of kilohertz and tens of kilohertz. For more details on frequency response tests on super-high-speed PMSGs, see [Reference 26](#).

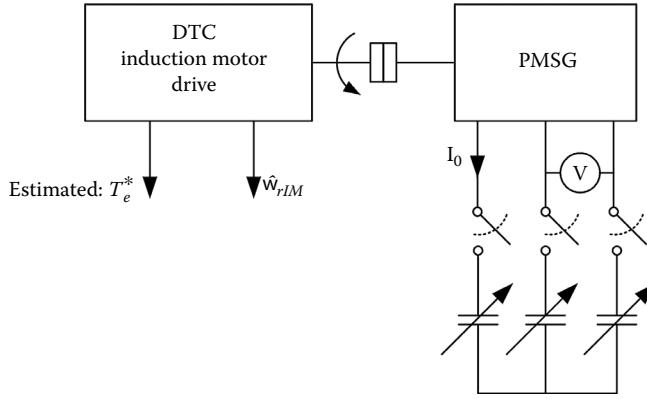


FIGURE 10.59 Generator no-load test arrangement.

10.15.2 No-Load Generator Tests

No-load generator tests need a small power prime mover. Today, variable-speed induction motor drives, with torque and speed estimation, are available off the shelf and may be used to drive the PMSG on no load at various speeds (Figure 10.59).

The no-load generator tests at various speeds may be used to identify the mechanical plus core losses at those speeds:

$$p_{mec} + p_{iron} \approx \hat{T}_e^c \cdot \frac{\hat{\omega}_{rmin}}{p_1} \quad (10.145)$$

with p_1 equal to the number of pole pairs in the induction machine (IM) driver.

$\hat{T}_e^c$ is corrected for mechanical loss torque of the IM. To do so, the IM drive is first driven uncoupled from the PMSG, and the torque is again estimated. The errors of this estimation are not small, however. Alternatively, the IM drive losses may be segregated in advance.

On no load, with IM electrical speed measured, the PMSG electric speed ω_r is as follows:

$$\omega_r = \omega_{rIM} \frac{(p_1)_{PMSG}}{(p_1)_{IM}} \quad (10.146)$$

The PM flux linkage Ψ_{PM} is

$$\Psi_{PM} = \frac{V_{0line(RMS)} \sqrt{2}}{\sqrt{3} \cdot \omega_r} \quad (10.147)$$

It is feasible to connect variable capacitors (Figure 10.59) to modify the magnetization state, mainly along axis d :

$$\omega_r \Psi_d = \omega_r (\Psi_{PM} + L_d I_d) = \frac{I_d}{\omega_r C}; \quad I_d = I_{os} \sqrt{2} \quad (10.148)$$

Consequently, the $\Psi_d(i_d)$ curves may be obtained with variable capacitors only.

The core losses also change due to i_d armature mmf presence, and this eventually is sensed by using, again, Equation 10.145.

The separation of p_{mec} from p_{iron} may not be done through no-load generator tests unless the capacitor is variable. But many times, this separation is not necessary.

10.15.3 Short-Circuit Generator Tests

The same arrangement as those for no-load generator tests may be used for short-circuit steady-state tests, to be operated at a few frequencies. Above a certain frequency,

$$I_{sc3} = \frac{I_d}{\sqrt{2}} \approx \frac{\omega_r \Psi_{PM}}{\sqrt{2} \omega_r L_{du}} = \frac{V_{0l(RMS)}}{\sqrt{3} \omega_r L_{du}} \quad (10.149)$$

L_{du} is the unsaturated value of d axis inductance.

The core losses are small in this test, but besides mechanical losses and stator-winding losses, stray-load losses in the rotor may be measured in this test if p_{mec} is already somehow known:

$$\sum p = p_{mec} + \frac{3}{2} R_s (1 + k_{skin}) I_d^2 + p_{stray}(\omega_r, I_s) \quad (10.150)$$

The rotor losses p_{stray} are produced by the space harmonics of the stator mmf augmented by the stator slot openings and by the time harmonics of the stator currents (mmf). It may be argued that we lumped into p_{stray} the stator surface losses due to airgap PM flux density harmonics caused by stator slot opening. This is true, but this is the core of the strayload losses definition.

Considering that the mechanical losses are known, and also the stator resistance at low frequency is known, the skin effect coefficient k_{skin} (in Equation 10.150) may be obtained through a test without the rotor in the airgap. If this test is performed at frequency f_1 up to maximum frequency with a PWM voltage source inverter, only the winding losses are important, and thus, with $(R_s)_{dc}$ known, current and power measured, $k_{skin}(f_1)$ may be determined.

10.15.4 Stator Leakage Inductance and Skin Effect

The test with the rotor absent in the airgap (Figure 10.60a) may also be used to separate the reactive power:

$$Q_3 = 3\omega_1 L_{sl3} I_{s0}^2 / 2 \quad (10.151)$$

The machine inductance is only marginally larger than the stator leakage inductance. One more test in the same situation, with all phases in series (Figure 10.60b), provides for the homopolar inductance L_{so} :

$$Q_{10} = 3\omega_1 L_{so} I_{s0}^2 / 2 \quad (10.152)$$

L_{so} is smaller than L_{sl} . Consequently, from the two tests, the stator leakage inductance L_{sl} may be safely calculated as an average of the two:

$$L_{sl} = (L_{sl3} + L_{slo}) / 2 \quad (10.153)$$

If the null point of windings is not available, it is possible to connect phase a in series with phase b and c in parallel, as in Figure 10.60c, and repeat the tests. Again, the inductance measured is around the leakage inductance. The reactive power is as follows:

$$Q_1 \approx \frac{3}{2} \omega_1 L_{slabc} I_{s0}^2 / 2 \quad (10.154)$$

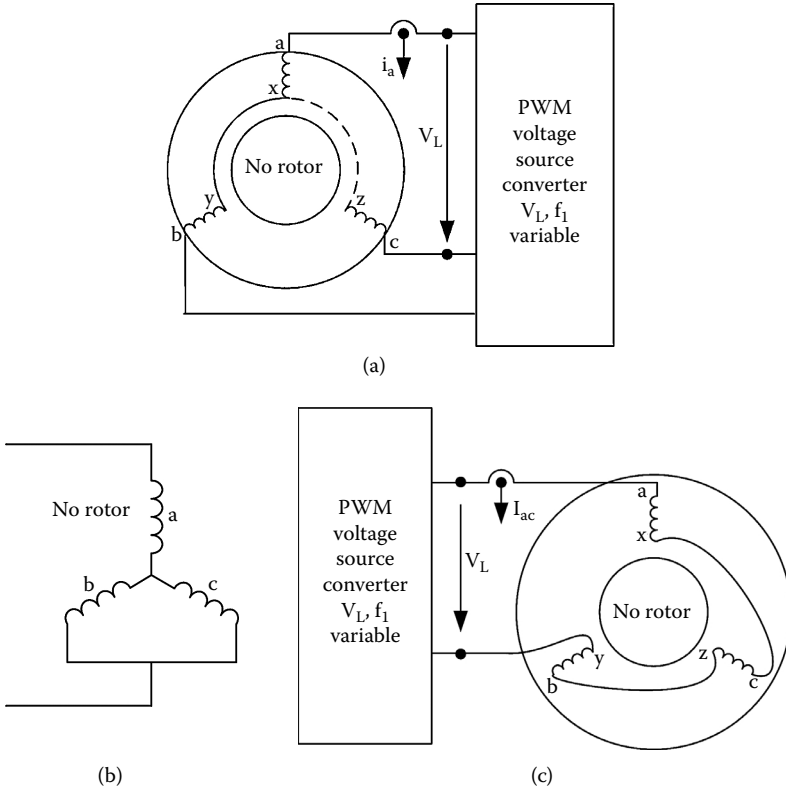


FIGURE 10.60 Alternating current (AC) tests without rotor in place: (a) three-phase, (b) a–bc connection, and (c) phases in series.

Notice that I_s is the current space-phasor amplitude $I_s = I_1 \cdot \sqrt{2}$, where I_1 is the stator-phase RMS current. If the stator current contains time harmonics, the leakage inductance has to be calculated by segregating and using only the current and reactive power fundamentals.

10.15.5 The Motor No-Load Test

The motor no-load test at variable voltage is standardly used to segregate mechanical losses from iron losses in motors. It may also be used for the PMSG, but if a notable stator mmf reaction occurs, the iron losses on load are notably larger than on no load.

Running the PMSG as a motor, even at no load (mechanical no load), implies using either a PWM converter with position-triggered control or driving it somehow near synchronism and with self-synchronization to the power grid.

The total losses (active power) in this test are as follows:

$$\sum p = p_{mec}(\omega_r) + p_{iron}(V_s^2) + \frac{3}{2} R_s I_{som}^2 \quad (10.155)$$

During the test for variable voltage V_s and the same speed (frequency, ω_r), p_{mec} remains constant, but p_{iron} varies with V_s^2 (if eddy current core losses are notably larger than hysteresis losses).

When representing Equation 10.155 on a graph (Figure 10.61), its intercept with the vertical axis represents the mechanical losses for the given speed ω_r .

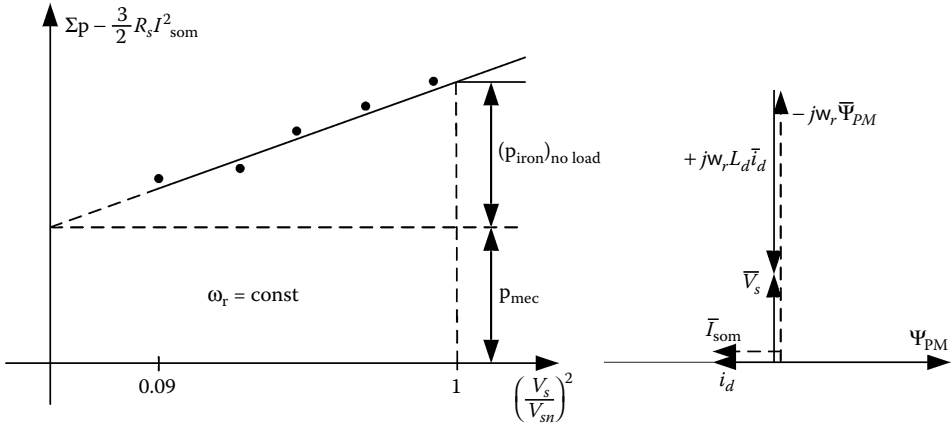


FIGURE 10.61 Segregation of mechanical losses in the no-load motoring test.

The no-load motoring test may also be used to approximately obtain the d axis (PM) magnetization curve as follows:

$$V_s \approx \omega_r \Psi_d(i_d); \quad i_d \approx i_{som} \quad (10.156)$$

Next, the machine current i_d changes sign. In general, at rated voltage $V_{sn} < V_0 = \omega_m \Psi_{PM}$, the reduction of V_{sn} during the test keeps the machine overexcited. Consequently, the armature reaction reduces the total flux in the machine (i_d stands for demagnetizing).

Note that generating for the same voltage and speed would mean higher total emf and, thus, heavier magnetic saturation. Operating the motor on no load at a higher than rated generator voltage by 3 to 4% might produce more realistic results.

10.15.6 The Generator Load Tests

The load tests should generally be performed for the same conditions as in the real operation mode. That is, if a diode or an active front-end rectifier is used in the application, it has to be there during the tests. However, it is also feasible, for novel configurations and proof of principle prototypes, to use a simplified test arrangement (Figure 10.62a and Figure 10.62b).

The variable capacitor bank is used to simulate PMSG voltage boosting to keep, for example, the load voltage constant at constant speed, when the load increases.

The on-load tests are used to verify temperatures and to determine the efficiency as the input P_{in}^{mec} and output P_{out}^{el} are measured:

$$\eta_{PMSG} = \frac{P_{out}^{el}}{P_{in}^{mec}} \quad (10.157)$$

High-precision power measurements are required to render the real value of efficiency.

The DC voltage vs. DC current $V_{dc}(I_{dc})$, for given speed and various capacitors or for various speeds and loads, may be obtained with diode rectifier load. For the AC load Z_L , with given R_L/L_L ratio, the generator voltage V_1 vs. phase current I_1 can also be obtained. $V_1(P_{out}^{el})$ can be obtained as in stand-alone tests (some results were shown in Section 10.6 and Section 10.7 of this chapter). The capacitor C_f on the DC side is replaced by a battery if this is the case in the application.

The load testing, basically with resistive inductive AC or DC load, may also be used to detect at least one (q axis) of the d - q magnetization curves if the other (d axis) is known from no-load tests.

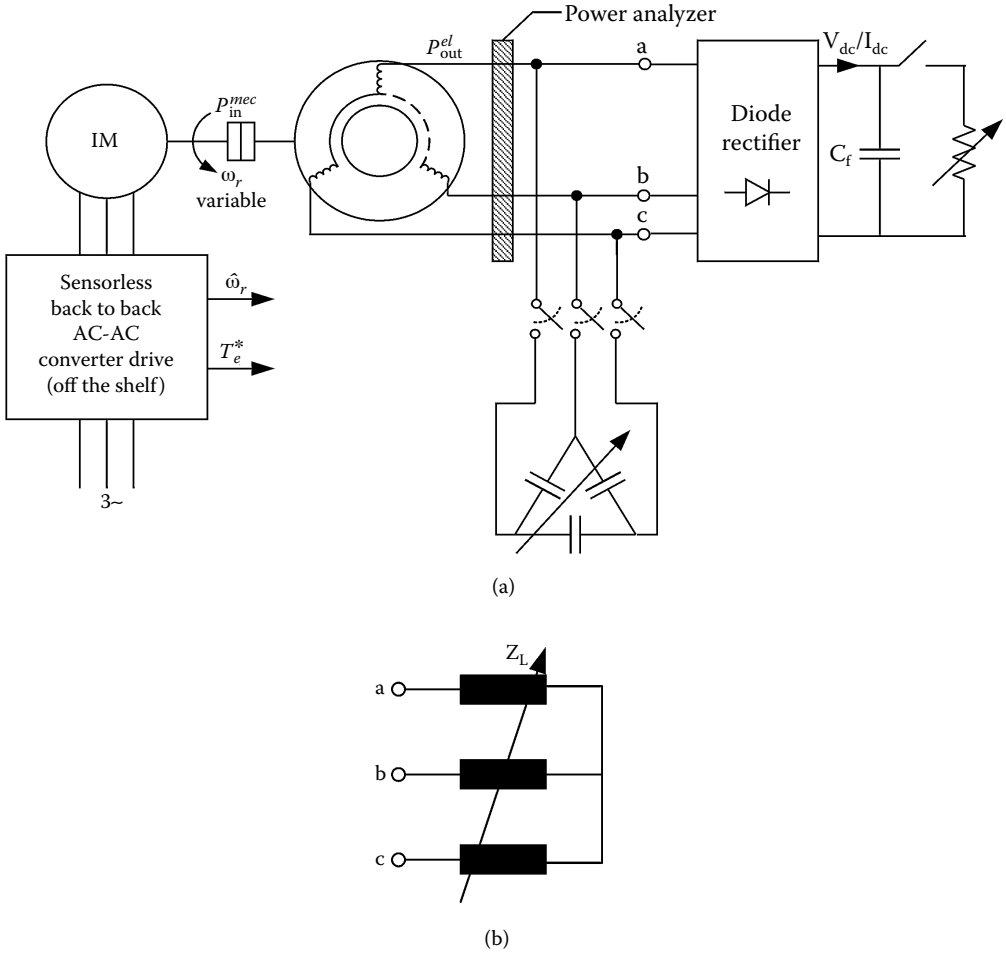


FIGURE 10.62 Load testing arrangements: (a) with diode rectifier load and (b) with alternating current (AC) load.

Let us consider the steady-state equations in the d - q model:

$$\begin{aligned} i_d R_s + V_d &= +\omega_r \Psi_q; \quad \Psi_q = L_q i_q \\ i_q R_s + V_q &= -\omega_r \Psi_d; \quad \Psi_d = \Psi_{PM} + L_d i_d \end{aligned} \quad (10.158)$$

From Equation 10.158 and Figure 10.63,

$$V_d = -V_s \cos \delta_V < 0; \quad V_q = -V_s \sin \delta_V < 0; \quad 0 < \delta_V < 90^\circ \quad (10.159)$$

$$\begin{aligned} I_d &= -I_s \sin(\delta_V + \phi_1) < 0; \quad I_s = I_1 \sqrt{2} \\ I_q &= -I_s \cos(\delta_V + \phi_1) < 0; \quad V_s = V_1 \sqrt{2} \end{aligned} \quad (10.160)$$

$$\begin{aligned} \Psi_d &= \frac{V_s \sin \delta_V + I_s R_s \cos(\delta_V + \phi_1)}{\omega_r} > 0 \\ \Psi_q &= \frac{-V_s \cos \delta_V - I_s R_s \sin(\delta_V + \phi_1)}{\omega_r} < 0 \end{aligned} \quad (10.161)$$

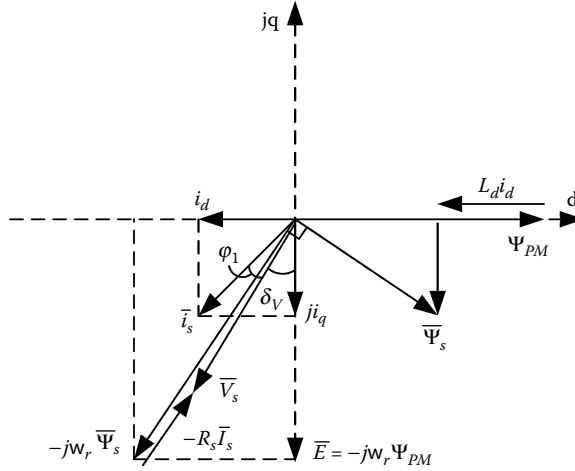


FIGURE 10.63 Vector diagram with permanent magnet synchronous generator (PMSG) delivering active and reactive power.

By measuring V_1 , I_1 and $\cos \phi_1$ (from P_1 , V_1 , I_1), Equation 10.161 still contains three unknowns Ψ_d , Ψ_q , and δ_V .

The zero power generator plus capacitor test produces the expression of Ψ_d :

$$\Psi_d = \Psi_{PM} + L_d i_d; \quad \sqrt{2} V_0 / \omega_{r0} = \Psi_{PM} \quad (10.162)$$

with L_d being basically constant.

The unsaturated inductance L_d may also be provided from the short-circuit test:

$$L_d = \frac{\omega_r \Psi_{PM}}{\omega_r I_{sc3} \sqrt{2}} = \frac{E}{\omega_r I_{sc3}} \quad (10.163)$$

E is the emf (PM-induced voltage) or the no-load voltage for given speed (frequency).

With Equation 10.162 in Equation 10.161, and L_d and Ψ_{PM} known,

$$\omega_r \Psi_{PM} + I_s (\omega_r L_d \sin(\delta_V + \phi_1) - R_s \cos(\delta_V + \phi_1)) = V_s \sin \delta_V \quad (10.164)$$

It is now obvious that at least iteratively, the power voltage angle δ_V may be calculated from Equation 10.164. In Reference 44, Equation 10.164 is manipulated toward a second-order algebraic equation in $x = \cos \delta_V$ ($\sin \delta_V = \sqrt{1 - x^2}$), whereby with the condition $\delta_V + \phi_1 < 90^\circ$, the solution is chosen. Once δ_V is known, from Equation 10.161, the magnetization curve along axis q , $\Psi_q(i_s)$, for various i_d , i_q combinations, may be obtained.

In a surface PM pole rotor, $L_d = L_q = L_s$, even if the machine gets saturated. Then,

$$\begin{aligned} \Psi_d &= \Psi_{PM} + L_s(i_d, i_q) i_d \\ \Psi_q &= L_s(i_d, i_q) i_q \end{aligned} \quad (10.165)$$

with Ψ_{PM} found from the no-load test, and after introducing Equation 10.165 in Equation 10.161, we obtain two equations with two unknowns — $L_s(i_d, i_q)$ and δ_V .

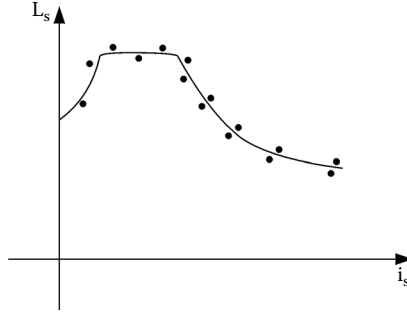


FIGURE 10.64 Unique inductance L_s vs. total current i_s for various (i_d, i_q) combinations.

A simple iterative solution of Equation 10.161, now under the form as follows,

$$I_s [\omega_r L_s \sin(\delta_V + \varphi_1) - R_s \cos(\delta_V + \varphi_1)] = V_s \sin \delta_V - \omega_r \Psi_{PM} \quad (10.166)$$

$$I_s [\omega_r L_s \cos(\delta_V + \varphi_1) - R_s \sin(\delta_V + \varphi_1)] = V_s \cos \delta_V \quad (10.167)$$

proves both variables $L_s(i_d, i_q)$ and δ_V . Then, it may be checked if the unique magnetization curve concept really applies for the case, by representing $L_s(I_s)$ with $i_s = \sqrt{i_d^2 + i_q^2}$, as in Figure 10.64.

If the experimental values of L_s calculated for various (i_d, i_q) do not fall around a unique $L_s(i_s)$ function, the family of curves $L_s(i_d, i_q)$ is to be used. The unique $L_s(i_s)$ function, if applicable, however, simplifies the digital simulations for transients and control design.

10.16 Note on Medium-Power Vehicular Electric Generator Systems

By medium-power vehicular electric generator systems, we mean here those systems with the powers in the hundreds of kilowatts to tens of MW used on aircraft and vessels. In terms of operation principles, they may be synchronous with excitation or PMs or induction or switched reluctance types.

All these types were treated in notable detail in previous chapters. This is why here we will dwell only on advanced configurations most suitable for aircraft, trains, and vessels.

The electric power on board aircraft is expected to reach in the future up to 500 kW per engine in a more electric aircraft.

Diesel engine long-haul or commuter trains with electric propulsion require better generators.

Electric propulsion is also typical in low dead weight vessel, and up to 20 to 30 MVA SGs are used as power sources to supply the electric propulsion motors through their power electronics control.

Up to 3 MVA generators with 690 V (line voltage, RMS value) output may be adopted.

Above these power levels, medium-output voltage generators up to 6 kV (line voltage, RMS value) are applicable.

Aircraft generators operate at speeds up to 30 to 60 krpm and powers up to 250 to 500 kW/unit. For diesel engine electric trains and vessel, the speeds are generally in the range of 1000 to 3000 rpm, and powers per unit from a few hundred kilowatts to MVA and tens of MVA.

Besides, starter generators on engine shafts at powers up to 250 to 500 kW/unit and emergency generators in the range of tens of kilowatts are also needed on aircraft. Three or four 250 kW generators have to be connected in parallel, or islanded, as required, for optimum use of energy and best reliability and power quality.

SGs with brushless excitation control for constant 400 Hz and constant voltage output are the standard on aircraft. Once full power electronics on board is accepted, PM or switched reluctance generators with the DC power bus at 270 V_{dc} for example [48–50], might come quickly.

SGs with excitation regulation for voltage control and speed regulation for frequency control are also standard on diesel engine electric trains and vessels.

However, at least up to 3 to 5 MVA multiple-pole PM generators with frequency up to 500 Hz, together with boost DC–DC converters to secure a constant DC voltage bus, may be adopted for trains and small vessel applications.

It should now be noticed that the above schemes were already discussed in the chapters on SG performance, design, and control, in this chapter and in [Chapter 9](#) on switched reluctance generators.

This is why here we do not pursue further the medium-power vehicular generator systems.

10.17 Summary

- By PMSG, we understand here a radial and axial airgap PM brushless generator with distributed or concentrated stator winding and surface PM or interior PM pole rotors.
- The PMSG applications span from wind turbine and microhydropower direct-driven electric generators, to automotive starter generators, or generators only for automotive applications, mobile or standby gensets with multiple output, and super-high-speed gas turbine generator systems for distributed power systems or for stand-alone applications.
- The market for PMSGs is gaining momentum worldwide, and the potential for fast growth seems very good.
- Rotors for PMSGs are built in cylindrical or disk shapes with surface PM or interior poles. Hybrid surface PM plus variable reluctance rotor sections may be combined for wide constant power speed range applications.
- For super-high speeds (above 15,000 rpm), a thin copper shield over the PMs and a carbon fiber resin retaining ring may be needed to reduce rotor electrical losses and provide for mechanical rigidity.
- Besides distributed windings, for powers up to 100 kW at 80,000 rpm, non-overlapping (tooth-embrace) coil windings may be applied, as they show high fundamental winding factors, lower copper losses, but slightly higher core losses as the number of poles is larger.
- The maximum fundamental frequency in super-high-speed PMSGs may go up to 2.5 kHz for up to 150 kW and up to 1 kHz for powers in the 1 to 5 MW range.
- For surface PM pole rotor PMSGs, in the absence of magnetic saturation, a comprehensive analytical model was built to explore steady state and transient performance. The space harmonics of the stator mmf airgap field with consideration of slot openings, through a permeance P.U. function, are included; thus, a realistic resultant airgap flux density distribution is built. A similar model was described for the axial airgap rotor PMSG, where an additional radial flux distribution change P.U. factor was introduced to make the model more realistic.
- Traditionally, core losses were calculated by analytic formulas under no-load conditions. It was recently shown that additional substantial core losses occur under load conditions. Analytical formulas for this case were also proposed but were only based on FEM-extracted time evolution of flux density in the finite element added contributions. Good correlation with test results was shown by such a combined method. The rotational character of the magnetic field in various machine core regions is very important to consider.
- For the investigation of transients, the phase coordinate model is first used. Then, the d – q model is shown to be a practical method for the study of transients and control performance and design.
- The d – q model may be used when the emf is sinusoidal and the phase inductances are either independent of or vary sinusoidally with rotor position. For the PMSG with shunt capacitors and AC load or diode rectifier load, the d – q model is instrumental in investigating the PMSG steady

state and transient performance. A potential oscillatory dynamics effect has to be eliminated in such schemes.

- When a PWM voltage source inverter is added to the diode rectifier, the PMSG system may deliver constant frequency and voltage power to a local power grid or to separate loads, at variable speed.
- It was proven that if a third harmonic in the stator emf of PMSG is produced, the latter may increase the output by as much as 15% with a three-phase diode rectifier for both series and bridge connections, with a loss of efficiency of about 1% only. The addition of the fourth diode leg to the three-phase bridge diode rectifier seems to pay off. A larger number of phases does not seem to produce better effects due to repetitive commutations.
- A variable inductance in parallel with a capacitor at machine terminals may produce very good dynamic performance in PMSG with constant speed controlled diesel engine prime movers and constant voltage output with load. A thyristor variac in series with the fixed inductances is controlled to regulate the load voltage at constant value. State feedback control systems perform well for such applications.
- When PMSGs have a diode rectifier at their terminals, the equivalent fundamental power factor angle in the machine φ_1 is forced to a small value: from zero at low frequencies to 10 to 15° at 1 kHz or so. In these conditions, the maximum torque, for $L_d = L_q = L_s$ (surface PM rotor), is as follows:

$$T_e = \frac{3}{2} p_1 \frac{\Psi_{PM}^2}{2L_s} \quad \text{or} \quad t_e(p.u.) = \frac{1}{2X_s(p.u.)}$$

This limitation may serve as a key design relationship, when sizing a PMSG of lower synchronous reactance $X_s(p.u.)$ is desirable.

- Gensets for mobile or standby power based on PMSGs may be built with multiple outputs, provided full power electronics control is provided. Variable speed is used to reduce fuel consumption. In such conditions, a DC voltage booster is required, besides the PWM voltage source inverter on the load side. As a 28 V_{dc} battery is used for the starter of the prime mover, the latter is used through a three-stage voltage booster to supplement the DC link voltage when it is too low momentarily (for a few seconds to minutes) until the speed of the engine recovers [24].
- The 50 (60), 400 Hz three- and single-phase output selection is a special asset of the new gensets, in addition to their reduced mass and volume in comparison with the standard ones that make use of voltage-controlled SGs at constant output frequency (and speed).
- PMSGs were recently proposed for automobiles at 42 V_{dc} to replace the existing claw-pole alternators, as more power is needed on board and at a higher efficiency (less fuel), with faster availability under load variations.
- PMSGs are being introduced for super-high-speed gas turbines for powers up to a few MW to reduce volume and cost, while providing fast active and reactive power control (at variable speed) on local power grids or separate loads.
- Mechanical constraints expressed in terms of maximum peripheral speed or rotor diameter vs. speed have to be observed for super-high-speed PMSGs.
- In super-high-speed PMSGs (rotor peripheral speed above 150 to 200 m/sec), the fundamental frequency may go up to 2.5 kHz for 100 to 200 kW and 1 kHz in the MW power/unity range. Higher frequency leads to a lower volume machine with lower copper losses but larger iron losses. It also causes large commutation losses in the PWM converter used for control.
- A super-high-speed rotor may run through one to three critical speeds before reaching the rated speed [46]. These critical speeds have to be driven through quickly, to avoid excessive noise and vibration.
- Super-high-speed PMSGs in the MW power range and 12,000 to 15,000 rpm require, in addition to the diode rectifier, with or without a DC link voltage booster, a PWM medium-voltage multilevel converter. Such converters are now available up to tens of MW and may also be used for the scope.

- Back-to-back two-level voltage source (six-leg) PWM converters [51] may be used for powers up to 1 MW or so to provide full four-quadrant operation with motoring, for starting the turbine.
- The electromagnetic design of a PMSG depends essentially on the operating modes, power, and output voltage range. Due to such variations, it seems practical to use a given rotor shear stress range (0.5 to 5) N/cm² to size the stator internal diameter D_{is} and the stack length for the maximum torque required in the design. From high-speed voltage constraints, the number of turns per phase is calculated after looking for the minimization of peak current for maximum torque at minimum speed. Winding tapping may be used for wide constant power speed range, to reduce the peak current for given output voltage; that is, to reduce the PWM converter costs.
- PMSGs are a novel technology, and thus, special standards for their testing are not yet available. Many of the standard tests (IEEE standard 115/1995) for regular synchronous machines may be applied to PMSGs as well. Standstill, no-load, and short-circuit generator, no-load motor, and load tests may all provide information on PMSG energy conversion performance (losses, efficiency, power at given voltage and speed) and parameter (Ψ_{PM} , L_d , L_q , R_s) estimation.
- Only mild cross-coupling saturation effects were found in PMSGs, and thus, they may be described by two unique magnetization curves along axes d and q : $\Psi_d^*(i_s)$, $\Psi_q^*(i_s)$ with the following inductances [47]:

$$L_d(i_d) = \frac{\left(\Psi_d^*(i_s) - \Psi_{PM}\right)}{i_s}; \quad L_q(i_s) = \frac{\Psi_q^*(i_s)}{i_s}$$

and

$$\Psi_d = L_d(i_s)i_d + \Psi_{PM}; \quad \Psi_q = L_q(i_s)i_q$$

where i_s is the total stator current $i_s = \sqrt{i_d^2 + i_q^2}$.

- The PMSG systems — with full power electronics digital control — may be considered a novel technology with exceptional growth potential in the near future for renewable energy conversion distributed power systems and for vehicular technologies [51].

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